FTD-MT-24-115-69-Vol II of II

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FOREIGN TECHNOLOGY DIVISION

AIR FORCE SYSTEMS COMMAND

PRINCIPLES OF JAMMING AND ELECTRONIC RECONNAISSANCE

by

S. A. Vakin and L. N. Shustov

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EDITED MACHINE TRANSLATION

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English Pages: 258 to 477

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FTD-MT- 24-115-69-Vol II of II Date 23 May 1969

CHAPTER 6

JAMMING OF RADIO CONTROL AND COMMUNICATION LINKS

6.1. Introduction

Suppression of command radio control and communication links permits solving the problem of RPD, inasmuch as this ensures description of the guidance circuits of fighters or guided anti-aircraft missiles [ZUR]. Actually, if the guidance circuit as its radio network includes a radio control link, its suppression leads to full breakdown of guidance, whereas in the case of radar jamming even full suppression of the latter does not always involve description of the circuit, since, in principle, it is possible to determine at least the angular coordinates of the source of interference and to carry out guidance to it.

Command radio control and communication links can be dealt two forms of active jamming:

- amplitude-modulated noise and random pulse interference;

- simulation jamming.

The second form of interference (simulation jamming) constitutes interference signals analogous to desired signals but carrying false information. Sometimes this interference is also called diversionary.

6.2. Equation for Jamming Radio Control and Communication Links

Let us determine the dependence of the ratio of noise power to signal power at the input of the receiver of the suppressed radio link on parameters of jamming stations and the suppressed device.

Let us consider the general case of jamming fighter (or missile) radio control links when the covered target aircraft Ц and the aircraft carrying the RPD equipment П are not coincident in space (Fig. 6.1). The object of jamming in this case is the radio receiver aboard the fighter (or missile).

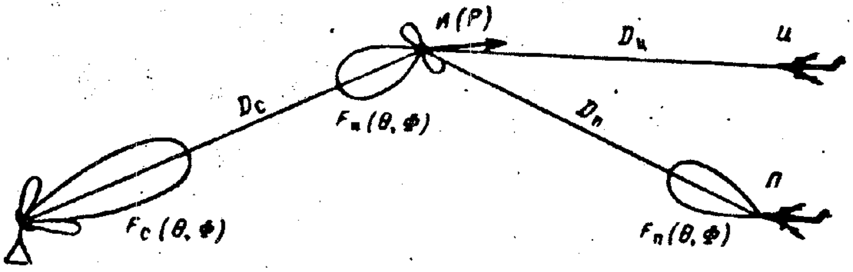

Fig. 6.1. Variant of jamming radio communication and control links.

Assuming that the peak of the radiation pattern of the receiving antenna coincides with the direction to the transmitter of the radio link, we will find the power of the desired signal (control signal) at the input of the receiver of the radio control link

$$P_{\text{с вх}} = \frac{P_c G_c}{4\pi D_c^2} A_r, \qquad (6.1)$$

where $P_c G_c$ – power potential of the radio control (or communications) transmitter;

A_r – maximum value of effective area of the receiving antenna on the missile (fighter);

D_c – distance between missile (fighter) and the transmitting station of the radio link.

The power of interference at the input of the receiver on the missile (fighter) is determined by relationship[1]

$$P_{\text{п вх}} = \frac{P_{\text{п}} G_{\text{п}}}{4\pi D_{\text{п}}^2} A_r F_{\text{и}}^2(\theta, \Phi) \gamma_{\text{п}} \frac{\Delta f_{\text{пр}}}{\Delta F_{\text{п}}}, \qquad (6.2)$$

where $P_{\text{п}} G_{\text{п}}$ – power potential of jamming transmitter;

$F_{\text{и}}(\theta, \Phi)$ – field pattern of receiving antenna of missile (fighter);

$\gamma_{\text{п}}$ – coefficient considering difference in polarizations of jamming transmitter antenna and receiving antenna of suppressed device;

$\Delta f_{\text{пр}}$ – bandpass of suppressed receiver of radio link;

$\Delta F_{\text{п}}$ – width of power spectrum of jamming transmitter;

$D_{\text{П}}$ – distance between missile (fighter) and jammer.

Using relationships (6.1) and (6.2), we find the sought equation for jamming

$$k = \left(\frac{P_{\text{п}}}{P_{\text{с}}}\right)_{\text{вх}} = \frac{P_{\text{п}} G_{\text{п}}}{P_{\text{э}} G_{\text{с}}} \left(\frac{D_{\text{с}}}{D_{\text{п}}}\right)^2 F_{\text{и}}^2(\theta, \Phi) \gamma_{\text{п}} \frac{\Delta f_{\text{пр}}}{\Delta F_{\text{п}}}. \qquad (6.3)$$

Let us assume that the suppression ratio for noise jamming of radio control links is equal to

$$k_{\text{п}} = 1.$$

Here by suppression ratio $к_{\text{П}}$ is understood the minimum necessary ratio of interference power to the power of the desired signal at the input of the suppressed receiver within the limits of the bandpass of its linear part, for which with given probability the possibility of reception of information by the suppressed receiver is excluded.

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[1] It is considered that the peak of the radiation pattern of the jamming transmitter antenna coincides with the direction to the fighter, i.e., $F_{\text{И}}(\theta, \Phi) = 1$.

Let us estimate the necessary power potential of the jamming transmitter for suppression of a fighter control link.

We will assume that at receiving and transmitting ends of the radio link and at the jamming station broad-beam antennas with identical polarization are used. Therefore it is possible tentatively to consider

$$G_{\text{с}} \approx G_{\text{п}} \approx 1, \ \gamma_{\text{п}} \approx 1, \\ F(\theta, \Phi) \approx 1. \tag{6.4}$$

We will assume also that selective interference with carrier frequency is created:

$$\Delta f_{\text{пр}} = \Delta F$$

In order to completely disrupt guidance of the fighter it is necessary to break the radio link up to the moment of locking of the homing guidance circuit (before detection and lock-on of the fighter's radar are accomplished). Thus the minimum jamming range will be at least equal to the uliimate range of the aircraft's radar, i.e.,

$$D_{\text{п мин}} = D_{\text{РЛС}}. \tag{6.5}$$

From (6.3), taking into account relationships (6.4) and (6.5), we obtain equation[1]

$$k = \frac{P_{\text{п}}}{P_{\text{с}}} \left(\frac{D_{\text{с}}}{D_{\text{РЛС}}} \right)^2. \tag{6.6}$$

whence, considering $k = k_{\text{п}} = 1$, we obtain the expression for determination of required power of jamming transmitter

$$P_{\text{п}} = P_{\text{с}} \left(\frac{D_{\text{РЛС}}}{D_{\text{с}}} \right)^2. \tag{6.7}$$

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[1] It is considered that the source of interferences is on the covered aircraft.

If we set D_c = 500 km, $D_{РЛС}$ = 50 km, $P_c = 10^4$ watts, then $P_{п\,мин}$ = = 100 watts.

We will now estimate the power potential of the noise jammer necessary for suppression of a missile (ZUR) radio control link, assuming the application of directional antennas.

We will set $k_п = 1$, $\Delta f_п = \Delta F_п$, $\gamma_п = 1$. Thus from (6.3)

$$P_п G_п = P_c G_c \left(\frac{D_п}{D_c}\right)^2 \frac{1}{F_в^2(\theta, \Phi)}. \tag{6.8}$$

Let us determine the minimum power potential $(P_п G_п)_{мин}$ necessary for suppression of ZUR control links. It is necessary to consider two cases of guidance: combined and command.

Combined Guidance

For the combined method guidance in the first stage is carried out by means of commands transmitted over a radio control link. In the last stage the ZUR is switched to conditions of homing guidance. In some cases the homing guidance phase can be excluded, especially under conditions of jamming of the homing device.

Minimum range of jamming transmitter $D_{п\,мин}$ is equal to ultimate range $D_{ГСН\,макс}$, at which the homing device (GSN) starts to function, i.e.,

$$D_{п\,мин} = D_{ГСН\,макс}.$$

For a rough estimate we will assume

$$\frac{D_п}{D_c} = 0{,}1.$$

It is necessary to consider that in examined case interference can be created only with respect to the side lobes of the antenna radiation patterns of guided antiaircraft missiles. Therefore it is necessary to take a value of $F^2_{\text{и}}(\theta,\Phi)$ of not over 0.01. Then with the help of (6.8) we obtain

$$(P_{\text{п}}G_{\text{п}})_{\text{мин}} = P_{\text{c}}G_{\text{c}},$$

i.e., the power potential of the jamming transmitter should be at least equal to the power potential of the station transmitting the commands.

Command Guidance

In command guidance the guidance circuit functions up to the moment of impact of ZUR and target. Therefore the minimum range of the RPD in this case should be selected on the basis of the necessity of providing assigned safe miss distance. It follows from this that conditions of suppression in command guidance are somewhat better than in combined guidance.

6.3. Forms of Jamming of Command Radio Control Link

The object of jamming in the KRU is the receiver aboard the interceptor or missile.[1] Active jamming can lead to suppression of transmitted commands or to creation of false commands, causing considerable errors in guidance.

In Fig. 6.2 is a simplified functional diagram of a single-channel KRU. The result of solution of the problem of guidance in the form of control command $K_{\text{вх}}$ is fed from the computer SRP to the line coder Ш. In the coder subcarrier signals are modulated so that they bear information on value and sign of control command $K_{\text{вх}}$. The subcarriers, in turn, modulate the carrier generated by the transmitter Prd.

The radiated signals are picked up by the receiver Prm on fighter or missile and after amplification and detection is fed to decoder DSh, which demodulates the subcarriers and forms the command $K_{\text{вых}}$, used as control signal (for example, in the automatic pilot).

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[1] Command radio control links and investigations of the influence of jamming on them are discussed in greater detail in works [38, 53].

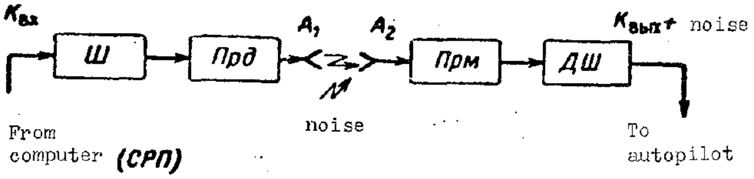

Fig. 6.2. Simplified block diagram of single-channel KRU.

In the KRU are applied different forms of modulation of subcarriers by signals taken from the SRP: amplitude, frequency, phase, pulse-width, pulse-code, etc. Jamming of the KRU regardless of the form of modulation applied in them, leads, as a rule, to the same results. Analysis of the influence of interference on the KRU with PPM is simplest.

In a radio link with PPM for transmission of command $K_{\text{вх}}$ phase modulation of pulsed subcarrier is used. The coder forms two groups of code pulses. The first group is the reference signal, the second the working signal. Each of these groups has its own time code. In Fig. 6.3a are represented pulse sequences with period T, consisting of two-pulse time codes (reference OK and actuating IK). The difference in time intervals T_1 and T_2 between last pulses of reference and actuating codes determines the value of the command coefficient

$$K_{\text{к вх}}=\frac{T_1-T_2}{T},$$

where $T = T_1 + T_2$ is command repetition period.

Signals radiated by the KRU transmitter, are picked up by the receiver aboard the fighter or missile and deciphered. The decoder contains a bistable multivibrator, which switches from one state to the other each time reference or actuating signals appear at its input.

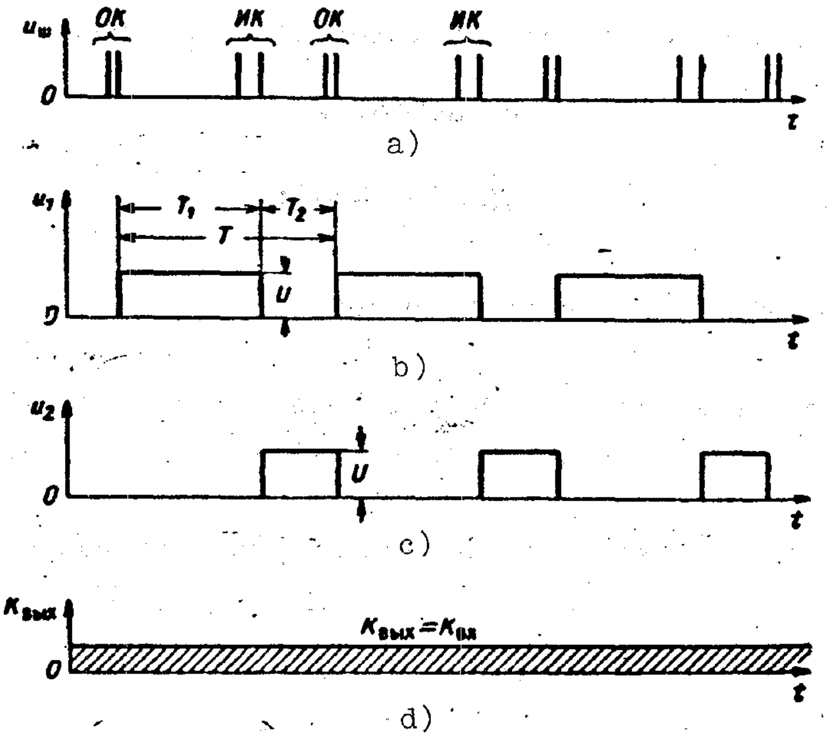

Fig. 6.3. Time diagrams, illustrating the operation of a single-channel KRU with pulse-phase modulation: a) transmitted sequences of reference (OK) and actuating (IK) codes; b) voltage at output of first tube of decoder trigger; c) voltage at output of second tube of decoder trigger; d) output control command.

If the multivibrator uses grounded-plate circuits, at the outputs of its first and second tubes (taking into account possible inclusion of phase-inverter stages) pulse voltages u_1 and u_2, depicted in Fig. 6.3b, c, are formed. The duration of output pulses of the trigger during normal operation (in the absence of interference) coincides with the duration of time intervals T_1 and T_2 between reference and actuating trains. The difference unit of the decoder forms the output control command[1]

$$K_{\text{вых}} = \bar{u}_1 - \bar{u}_2, \tag{6.9}$$

where $\bar{u}_1$ and $\bar{u}_2$ – time-averaged voltages u_1 and u_2, so that equality of input and output command coefficients is observed

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[1] For simplicity it is considered that the transfer functions of separate sections of the receiver (filter, subtractor, etc.) are equal to unity.

$$K_{\text{к вых}} = K_{\text{к вх}}, \tag{6.10}$$

where

$$K_{\text{к вых}} = \frac{\bar{\tau}_1 - \bar{\tau}_2}{U};$$

U — amplitude of trigger output pulses.

Interference acting on the KRU receiver distorts the transmitted commands, in consequence of which equality (6.10) is disturbed.

We distinguish two forms of jamming command radio links:

- code-barrage;

- selective-code.

Interference signals of the first form constitute continuous noise oscillations or random pulse interference (KhIP). In the second case the interference signals have structure similar to the useful signals and are created, as a rule, by relaying the desired signal with appropriate treatment.

Code-barrage jamming, in the form of continuous noise signals or KhIP, can cause:

- full or partial suppression of transmitted commands,

- change of parameters of modulation of subcarrier oscillations (for example, random deviations of moments of appearance of pulses at the trigger input with respect to the middle of operating pulses, and others),

- formation of false trains.

Depending upon the intensity of interference signals, we distinguish interference of low and high levels. Interferences of low level causes only slight changes of parameters of subcarrier modulation, which cannot lead to noticeable guidance errors. Interference of high level, for which $P_п/P_с \gg 1$, causes not only distortion of commands at the KRU output but also impairment of characteristics of the KRU, in particular, reduction of transfer function. When $P_п/P_с \geqslant k_п$, where $k_П$ is the suppression ratio of the given KRU, the transfer function of the radio link is reduced so much that it, as the dynamic link of the guidance circuit, is broken. Subsequently we will examine the influence on the KRU of interference of high level only.

During the influence of interferences on the KRU with pulse-position modulation there can be suppression of reference and actuating codes and formation of false codes, due both to noise pulses and combinations of noise and operating pulses (Fig. 6.4a). As a result there will be false triggering of the trigger (Fig. 6.4b and c) or, as we sometimes say, "splitting" of output pulses occurs. At the outputs of first and second tubes of the trigger are formed random pulses with identical amplitude U but with different (random) repetition period and duration. The mean value of the output command coefficient $\overline{K}_{к\,вых}$ in this case equals

$$\overline{K}_{к\,вых} = \frac{\overline{K}^{(п)}_{вых}}{(K_{вых})_{макс}} = \frac{\overline{u}_1^{(п)} - \overline{u}_2^{(п)}}{U}, \tag{6.11}$$

where $\overline{K}^{(п)}_{вых}$ – mean value of output command; $\overline{u}_1^{(п)}$ and $\overline{u}_2^{(п)}$ – mean value of output voltages of first and second tubes of the trigger in the presence of interference (Fig. 6.4b and c); U – amplitude of output voltages of trigger tubes, the trigger being considered symmetric.

Mean values of output voltages $\overline{u}_1^{(п)}$ and $\overline{u}_2^{(п)}$ can be found by the formulas

$$\overline{u}_1^{(п)} = UP_1, \tag{6.12}$$

$$\overline{u}_2^{(п)} = UP_2, \tag{6.13}$$

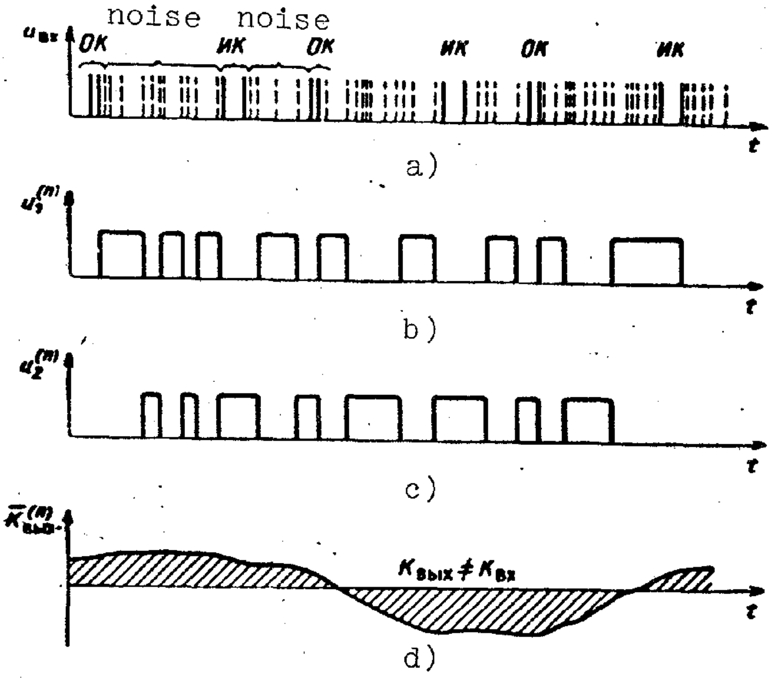

Fig. 6.4. Time diagrams, illustrating the action of code-barrage jamming on the KRU with PPM: a) pulses of reference (OK) and actuating (IK) codes and interference at decoder input; b) voltage at output of first tube of decoder trigger; c) voltage at output of second tube of decoder trigger; d) output control command.

where P_1 and P_2 are average probabilities that trigger output voltage $u_1^{(п)}$ and $u_2^{(п)}$ will be equal to U.

Probabilities P_1 and P_2 are determined by the form of interference and, in general their calculation is difficult. However, with accuracy sufficient for practice in a number of cases one may assume that moments of appearance of noise pulses at both trigger inputs obey Poisson law. Then the following expressions are valid [38]:

$$P_1 = \frac{1}{2T\lambda} e^{-2\lambda T_2}, \qquad (6.14)$$

$$P_2 = \frac{1}{2T\lambda} e^{-2\lambda T_1}, \qquad (6.15)$$

where λ is average number of pulses formed at both trigger inputs in 1 s.

Substituting expressions (6.12), (6.13), (6.14), and (6.15) in formula (6.11), after making simple transpositions, we obtain

$$\overline{K}_{\text{к вых}} = \frac{1}{2T\lambda}\left[e^{-\lambda T(1-K_{\text{к вх}})} - e^{-\lambda T(1+K_{\text{к вх}})}\right]. \qquad (6.16)$$

In Fig. 6.5 is represented dependence of mean value of output command coefficient $\overline{K}_{\text{к вых}}$ on parameter λT, plotted in accordance with formula (6.16) for values of input command coefficient $K_{\text{к вх}}$ = 0.8, 0.5, and 0.2. The given dependences and formula (6.16) show that in the presence of interference the KRU becomes nonlinear (with respect to input command $K_{\text{вх}}$), with transfer function decreasing with growth of intensity of interference λ.

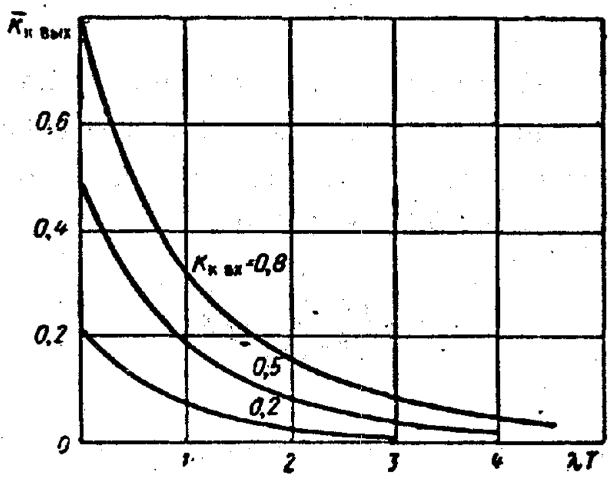

Fig. 6.5. Dependence of mean value of command coefficient $\overline{K}_{\text{к вых}}$ on parameter λT, characterizing intensity of interference.

Fluctuations of output command coefficient $K_{\text{к вых}}$ with respect to mean value $\overline{K}_{\text{к вых}}$ during the influence of interference of high level cannot play an essential role in appraisal of the effectiveness of jamming, since variance of output command coefficient $K_{\text{к вых}}$, in any case is one order of magnitude less than the mathematical expectation of $\overline{K}_{\text{к вых}}$.

The value of the transfer function of the KRU, as it is known, is determined by the formula

$$\kappa_{\text{КРУ}} = \frac{dK_{\text{к вых}}}{dK_{\text{к вх}}} \qquad (6.17)$$

with

$$K_{\text{к вх}} = 0.$$

Using relationship (6.17), from (6.16) we find

$$\kappa_{\text{КРУ}} = e^{-\lambda T}. \qquad (6.18)$$

During appraisal of the influence on the KRU of code-barrage interference the most general of its characteristics is the average number (z) of pulses of interference acting on the input of coincidence stages of the decoder in 1 s.

If at the KRU receiver input interference and desired signals are square pulses of identical duration and amplitude, there is a single-valued functional dependence of the value of z on the suppression ratio at the receiver input

$$z = \left(\frac{P_{\text{п}}}{P_{\text{с}}}\right)_{\text{вх}} mnF. \qquad (6.19)$$

Here $\left(\frac{P_{\text{п}}}{P_{\text{с}}}\right)_{\text{вх}}$ – ratio of average power of interference to average power of signal at the KRU receiver input;

m – number of pulse trains in one period T;

n – number of impulses in code train;

F = 1/T – repetition rate of trains.

The average number of pulses λ formed in a unit of time at both trigger inputs of the decoder is related to the value of z as[1]

$$\lambda = zP_{\text{к}} = \left(\frac{P_{\text{п}}}{P_{\text{с}}}\right)_{\text{вх}} mnFP_{\text{к}}. \qquad (6.20)$$

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[1] Formula (6.20) is valid if reference and actuating codes contain an identical number of pulses.

Here $P_{к}$ is the probability of jamming pulses combining to form reference or actuating code [54]:

$$P_{к} = n(z\tau_{и})^{n-1} = n^{n}\left[\left(\frac{P_{п}}{P_{с}}\right)_{вх} mF\tau_{и}\right]^{n-1}. \qquad (6.21)$$

For the examined KRU with PPM m = n = 2 (Fig. 6.3). Therefore, taking into account (6.20) and (6.21), we record formula (6.18) in the form

$$K_{КРУ} = e^{-32\left(\frac{P_{п}}{P_{с}}\right)_{вх} F\tau_{и}}. \qquad (6.22)$$

In Fig. 6.6 is a graph of dependence of the transfer function of the KRU ($K_{КРУ}$) on $\left(\frac{P_{п}}{P_{с}}\right)_{вх}$ for F = 100 Hz, $\tau_{и} = 10^{-6}$ s.

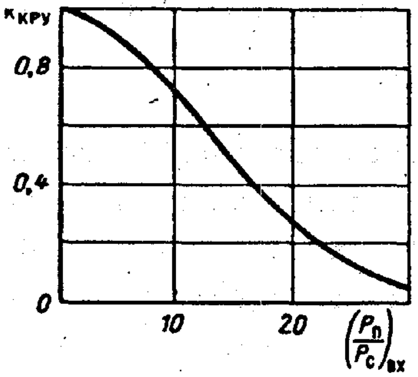

Fig. 6.6. Dependence of the standardized transfer function of the KRU on the suppression ratio.

This dependence shows that at the limit, when $\left(\frac{P_{п}}{P_{с}}\right)_{вх} \to \infty$, the transfer function of the KRU becomes equal to zero, which is equivalent to disruption of the link, as, consequently, is the entire guidance circuit. For practical purposes disruption of the link occurs with finite value of $\left(\frac{P_{п}}{P_{с}}\right)_{вх}$. The link can be considered open if its transfer function under the action of interference is decreased by 5-10 times. Hence, using Fig. 6.6, we find the ratio of average power of interference to average power of signals required for suppression of the examined KRU

$$\zeta \approx 25.$$

An increase of valency of code involves growth of ratio ζ.

Let us note that in accordance with accepted definition (1.3) the ζ is not the suppression ratio. The latter can be obtained if in the ratio $\left(\frac{P_п}{P_c}\right)_{вх}$ by P_c we understand pulse power of desired signal.

Depending upon the form of interference, construction of the KRU, and structure of the code, the influence of interference can lead to two qualitatively different cases of manifestation of the effect of disruption of the guidance circuit. Thus with a symmetric link, equally resistant reference and working codes, and interference in the form of KhIP disruption of the link means the setting of controls in neutral position.

Let us note that codes are equally resistant if during the influence of signal and interference the average numbers of interference pulses formed at inputs of first and second tubes of the trigger in a unit of time are equal. The KRU is considered symmetric if under conditions of interference the mean value of voltage at the output of the link is equal to zero (with zero command).

If, however, the link is asymmetric or unequal-resistance codes are applied, upon disruption of the link the controls of the missile will move to extreme positions, causing it to fly on a curvilinear trajectory with maximum permissible overload.

A significant deficiency of code-barrage jamming is the considerable value of necessary average power of jamming transmitter. This limits the practical application of such jamming.

Selective-code jamming, in principle, permits obtaining a certain power gain as compared to barrage jamming.

The principle of creation of such interference can be clarified with the aid of Fig. 6.7. Control signals (reference and actuating codes, Fig. 6.8) of the command radio control link are picked up by the receiving antenna of the jamming station $A_{П1}$ and fed to its receiver Prm and to the frequency memorization circuit SZCh. After amplification and detection signals of reference and actuating codes (Fig. 6.8a) are fed to the input of control unit UU, which converts these codes so that for each train or reference code one actuating code is formed and, conversely, instead of the actuating code a reference code is formed (Fig. 6.8b). The obtained sequence of pulses is delayed for time τ_3 and is then used for modulation of the high-frequency oscillations in amplifier У. After amplification in the final amplifier OU the interference signal is radiated by the transmitting antenna $A_{П2}$.

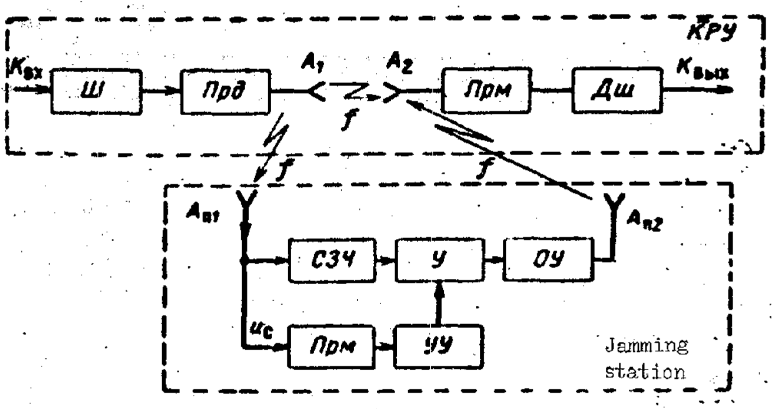

Fig. 6.7. Functional diagram, explaining the principle of creation of selective-code jamming of the KRU.

As a result of the influence of interference false triggering of the decoder trigger takes place. In Fig. 6.8c and d are depicted voltages at outputs of first and second tubes of trigger resulting from the influence of interference. Broken lines and shading denoted pulses formed by the trigger in the absence of interference.

As follows from the figure, in this case the output command $K^{(п)}_{вых}$ is distorted as compared to command $K_{вых}$, corresponding to operation without interference.

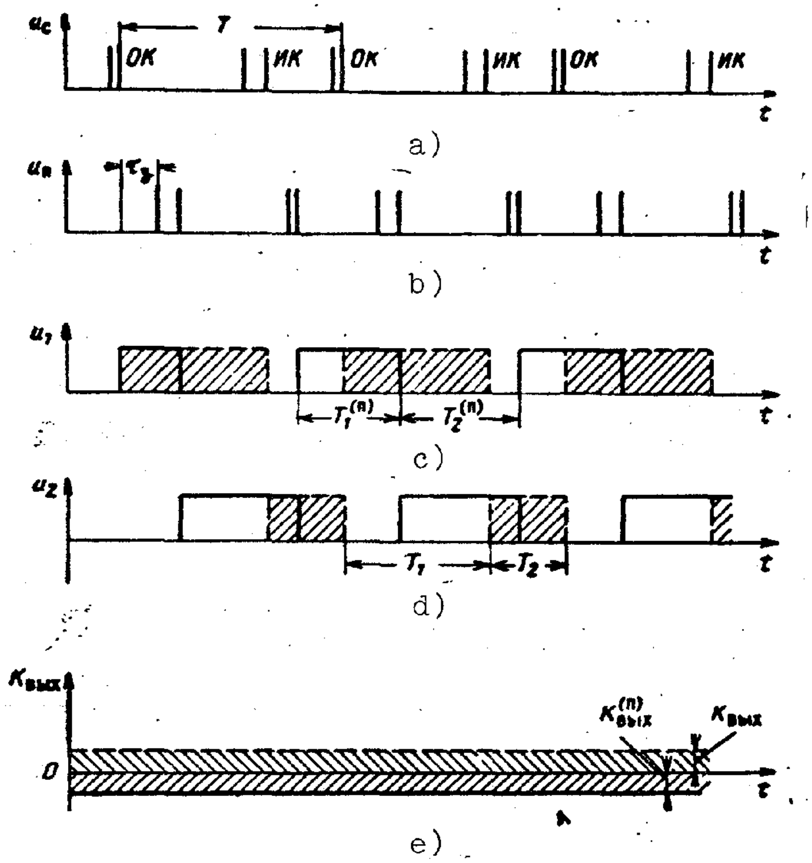

Fig. 6.8. Time diagrams, illustrating the operation of KRU with PPM under conditions of selective-code jamming: a) desired signals at input of KRU decoder; b) interference pulses at decoder input; c) voltage at output of first tube of decoder trigger; d) voltage at output of second decoder tube; e) output control command.

Necessary conditions for creation of selective-code interference are:

- reconnaissance of codes, their parameters, order and period of repetition;

- formation of similar interference codes and their radiation in corresponding order.

6.4. Forms of Jamming Radiotelephone Communications

The basic characteristic of the quality of communications is the intelligibility of speech (articulation), by which is understood the relative quantity of correctly received elements of speech from the total quantity of elements transmitted. Since speech is a random process, it makes sense to talk only about its statistical characteristics.

One of the important characteristics of speech is its formant spectrum, obtained as a result of statistical treatment of a great number of realizations. In Fig. 6.9 is shown the energy spectrum of Russian speech [55]. As follows from the figure, the biggest part of the energy of speech is contained in a rather narrow frequency band from 0 to 1000 Hz. However, the intelligibility of speech is considerably influenced even by those spectral components whose energy is low. It has been determined that the most important for intelligibility are the components of speech spectrum lying in the frequency band of 400-800 Hz. The intelligibility of speech usually is determined experimentally with the help of articulation tables. Under conditions of interference the intelligibility of speech, naturally, worsens, and at some threshold value of the ratio of interference power to signal power at the input of a receiver within the limits of its passband disturbance of radio communications sets in, i.e., the operator at the receiving end ceases to understand the meaning of transmitted information.

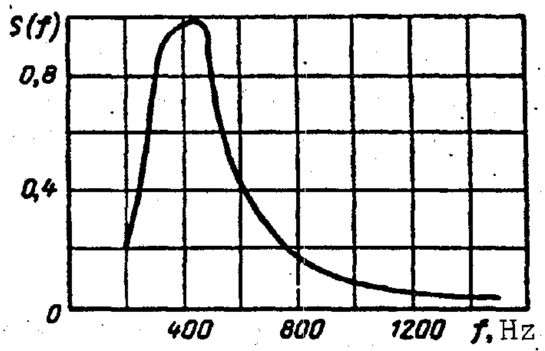

Fig. 6.9. Energy spectrum of Russian speech.

The ratio $k_п = \left(\frac{P_п}{P_c}\right)_{вх}$ is called the suppression ratio of radio communications link.

The value of the suppression ratio of a radio link is determined by the form of interference and its spectral characteristics, detuning of interference relative to the resonant frequency of the suppressed receiver, and the type of modulation applied.

The following forms of jamming can be used on radiotelephone communication links:

- amplitude-modulated noise;
- frequency-modulated noise;
- random-pulse;
- straight-noise.

As an example we will examine the influence on a receiver of amplitude-modulated interference.

At the receiver input we have

$$u_{\text{п}}(t)=U_{\text{п}}[1+m_{\text{п}}(t)]\cos(\omega_{\text{п}}t+\varphi_{\text{п}}), \qquad (6.23)$$

$$u_{\text{c}}(t)=U_{\text{c}}[1+m_{\text{c}}(t)]\cos(\omega_{\text{c}}t+\varphi_{\text{c}}), \qquad (6.24)$$

where $u_{\text{п}}(t)$ and $u_c(t)$ – instantaneous values of voltage of interference and signal; U_Π and U_c – amplitudes of interference and signal; $m_\Pi(t)$ and $m_c(t)$ – modulating functions of interference and signal; ω_Π and ω_c, ϕ_Π and ϕ_c – carrier frequencies and phases of interference and signal.

The envelope of combined voltage of interference and signal at the detector input can be recorded in the form

$$U_{\text{ог}}=K_{\text{ПРМ}}\sqrt{U^2_{\text{оп}}+U^2_{\text{ос}}+2U_{\text{оп}}U_{\text{ос}}\cos(\omega_{\text{п}}-\omega_{\text{c}})t}=$$

$$=K_{\text{ПРМ}}\sqrt{U^2_{\text{оп}}+U^2_{\text{ос}}}\sqrt{1+\frac{2U_{\text{оп}}U_{\text{ос}}\cos(\omega_{\text{п}}-\omega_{\text{c}})t}{U^2_{\text{оп}}+U^2_{\text{ос}}}}, \qquad (6.25)$$

where $U_{\text{о п}}=U_{\text{п}}[1+m_{\text{п}}(t)]$ – envelope of interference; $U_{\text{о с}}=U_{\text{с}}[1+m_{\text{с}}(t)]$ – signal envelope; $K_{\text{прм}}$ – gain of receiver (below for simplicity we take $K_{\text{прм}}=1$).

Expanding expression (6.25) in binomial series and limiting such to three members, which is proper when $m_{\text{с}}<1$ and $m_{\text{п}}<1$, as a result of simple transpositions we obtain the expression for detector output voltage

$$U=U_{=}+U_{\Omega\text{с}}+U_{\Omega\text{п}}+U_{\Omega\text{б}}, \tag{6.26}$$

where

$$U_{=}=U_{\text{о с}}\sqrt{1+b^2}; \tag{6.27}$$

$$U_{\Omega\text{с}}=\frac{U_{\text{с}}(2+b^2)}{2(1+b^2)^{3/2}}m_{\text{с}}; \tag{6.28}$$

$$U_{\Omega\text{п}}=\frac{U_{\text{п}}b(2b^2+1)}{2(1+b^2)^{3/2}}m_{\text{п}}; \tag{6.29}$$

$$U_{\Omega\text{б}}=\frac{U_{\text{п}}}{\sqrt{1+b^2}}\left[1+\frac{b^2}{1+b^2}m_{\text{с}}+\frac{1}{1+b^2}m_{\text{п}}\right]\cos\Omega_{\text{б}}t; \tag{6.30}$$

$$b=\frac{U_{\text{п}}}{U_{\text{с}}}>1;\ \Omega_{\text{б}}=\omega_{\text{п}}-\omega_{\text{с}}.$$

Expression (6.26) shows that disturbance of radio communications is due to the camouflaging action of the signal by the interference, which is provided by components of interference (6.29) and beats (6.30). Furthermore, there is direct suppression of the signal by noise when b > 1 in a nonlinear device, in this case the detector. Thus from (6.29) and (6.28) we have

$$\left(\frac{U_{\Omega\text{с}}}{U_{\Omega\text{п}}}\right)_{\text{вых}}=\frac{b^2+2}{2b^2+1}\,\frac{m_{\text{с}}}{m_{\text{п}}}\,\frac{1}{b^2}.$$

For example, when b = 2 and $m_{\text{с}}=m_{\text{п}}$

$$\left(\frac{U_{\Omega\text{с}}}{U_{\Omega\text{п}}}\right)_{\text{вых}}=\frac{1}{6}.$$

Consequently, for the conditions the suppression ratio at the detector output increases by 3 times.

In Fig. 6.10a is depicted the initial spectrum of interference and signal at the input of the suppressed device, while Fig. 6.10b shows the resultant spectrum of the signal at the detector output. From the figure it follows that with sufficient detuning $\Omega_{б}=\omega_{c}-\omega_{п}$ camouflaging of signal due to beats is eliminated.

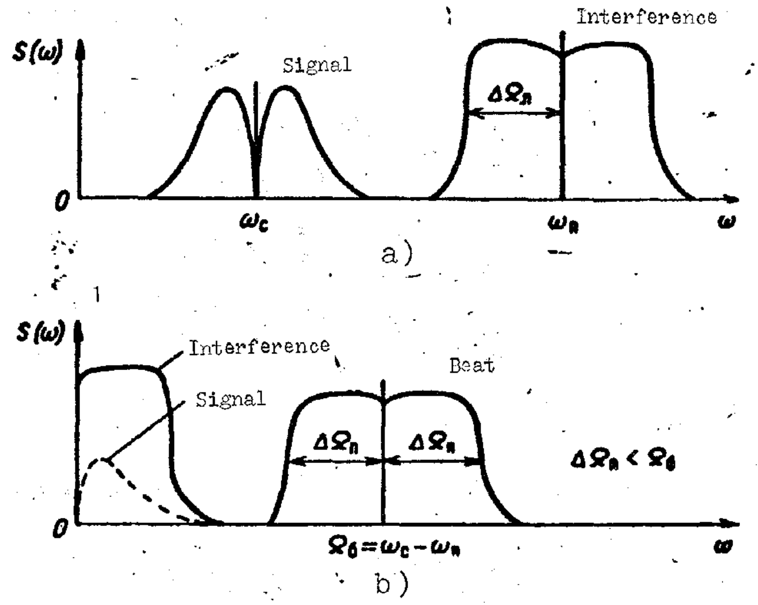

Fig. 6.10. Illustration of the camouflaging and suppressing effect of amplitude-modulated interference on radiotelephone communications: a) initial spectrum of interference and signal at input of suppressed receiver; b) resultant spectrum at detector output.

Therefore in the process of creation of spot jamming for radio links it is necessary to satisfy rather stringent requirements with respect to the accuracy of tuning of the amplitude-modulated interference to the carrier frequency of the signal. Analogous conclusions can be drawn with respect to other forms of jamming.

In Fig. 6.11 are presented experimental dependences of the suppression ratio $k_{П}$ of a communications receiver with amplitude modulation on the value of detuning $|\Delta f_{p}/\Delta f_{п}|$ of interference relatively to the resonant frequency of the suppressed receiver. These dependences were obtained by A. I. Velichkin for different forms of

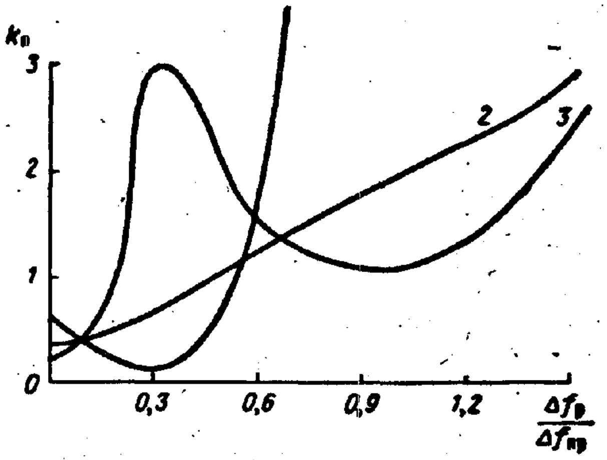

Fig. 6.11. Experimental dependences of the suppression ratio of the receiver of a communications link with amplitude modulation on detuning of interference relative to the resonant frequency of the suppressed receiver: 1 – random-pulse interference; 2 – combined interference; 3 – frequency-modulated noise interference.

interference: random-pulse (KhIP), frequency-modulated noise, and combined (amplitude-frequency modulated).

It is necessary to note that in this case dependence of suppression ratio on detuning exists because the width of the spectrum of spot jamming can be less than the passband of the suppressed receiver $\Delta f_{пр}$.

CHAPTER 7

PASSIVE RADIO INTERFERENCE

7.1. Introduction

By passive interferences are understood signals forming on the input of suppressed radio electronic devices as a result of the scattering of electromagnetic waves by objects employed in mass quantities. As a rule, there occurs the scattering of electromagnetic waves generated by transmitting antennas of suppressed radio electronic systems.

Sometimes passive interferences include decoys [false targets] and agents providing local ionization of space. It is expedient to distinguish these agents as independent ones. The basis for this classification are the following considerations.

Decoys are used in quantities measured in tens and hundreds of units. As a result of their application the electrical properties of the medium are not changed. With respect to the parameters of the reflected signal and the parameters of motion they should be identical to real targets.

An interference signal on a decoy (radar trap) is formed due to both passive re-emitters, and also active relays. Furthermore, on decoys (traps) special jamming transmitters are frequently mounted. All of this gives a basis to examine decoys (traps) as an independent form of RPD.

The application of agents ensuring the ionization of local regions of space as well as of various antiradar coverings leads to a change in the electrical properties of the medium, i.e., these means, by their very nature, in principle must exclude the possibility of using radio waves for purposes of measuring and transmitting information.

Usual passive interferences being re-emitters of a type of dipole reflectors are applied on a large scale; however, as a rule, a cloud of dipoles does not change the electrical properties of the medium, inasmuch as the distance between the dipoles in the cloud is many tens and hundreds of times greater than the wave length.

Therefore the action of passive interferences results in the formation of a camouflaging background and in this sense they are analogous to noise interferences.

Thus, it is necessary to distinguish three independent forms of passive agents of RPD, not requiring the use of jamming transmitters:

- decoys;
- passive interferences,
- agents changing the electrical properties of the medium.

At the present time passive interferences are created mainly by antiradar (dipole) reflectors ejected in great quantities into the atmosphere.

7.2. Dipole Reflectors

Dipole reflectors (dipoles) are made from paper, fiberglass, caprone covered with a conducting layer. It is possible to use metallic foil for this purpose. The length of the dipoles and their thickness are selected in such a way as to ensure the effective scattering of radio waves in as wide a frequency range as possible. As a rule, their length is approximately equal to half of the wave length of the suppressed radar. However dipoles are also used, the length of which considerably exceeds the wave length of the radar.

Dipole reflectors are usually put up in packs. Opening after ejection from the aircraft, such a pack creates a cloud of dipole reflectors, the reflected signal from which is observed on the screen of a plan position indicator (PPI) in the form of a bright spot. If one were to drop a rather large number of packs one after another, then on the PPI there would form illuminated bands of considerable lengths (Fig. 1.2b).

At the present time dipole reflectors are made on a dielectric basis or from foil. The minimum thickness of the metallic covering is determined by the thickness of the working surface layer formed due to the skin effect. The depth of penetration of the current into the conducting layer depends on the frequency of the electromagnetic oscillations. In the centimeter range the depth of penetration can be very small ($d \approx 1$ μm). This makes it possible to make the dipoles in the form of very thin strips of fibers with a diameter of several tens of microns. In practice it is necessary to consider the questions of strength and manufacturing technology.

The number of dipoles in a pack depending upon the range is tens of thousands and millions of units. On account of the incoherence of the fields, dispersed by the individual dipoles, the effective scatter area [ESA] of a cloud of reflectors of identical length will be on the average equal to the sum of the ESA of each dipole, i.e.,

$$\bar{\sigma}_{\text{п}} = \sum_{i=1}^{N} \bar{\sigma}_i = N\bar{\sigma}_1, \qquad (7.1)$$

where $\bar{\sigma}_{\text{п}}$ – the average ESA of a cloud of dipoles; $\bar{\sigma}_i = \bar{\sigma}_1$ – the average ESA of one dipole; N – the number of dipoles in a pack.

Formula (7.1) is correct in the ideal case, when all dipoles are employed effectively. In practice due to the entanglement (adhesion) of the dipoles and their breakage the ESA of a cloud will be less than $\bar{\sigma}_{\text{п}}$ determined by formula (7.1). Usually the ESA of a cloud of dipole reflectors is calculated by the formula considering the actual number of effective dipoles in a pack:

$$\bar{\sigma}_{\text{п}} = \eta N \bar{\sigma}_1, \tag{7.2}$$

where η – the coefficient of the effective number of dipoles.

The value of the effective scatter area (ESA) of one half-wave dipole ($\bar{\sigma}_1$) in general depends on its orientation relative to the electrical vector of the incident wave. Due to atmospheric turbulence and aerodynamic properties the dipoles are randomly oriented in the cloud, as a rule, relative to each other. Moreover, to safeguard the equiprobable orientation in the manufacture of dipoles the aim is that the center of gravity of each reflector be displaced at a random distance from its center.

Therefore the ESA of the whole cloud ($\bar{\sigma}_{\text{п}}$) is determined from the mean value of the ESA of one dipole ($\bar{\sigma}_1$) oriented randomly in space. the value of $\bar{\sigma}_1$ will be found below.

7.3. The Effective Scatter Area [ESA] of a Half-Wave Dipole Randomly Oriented in Space

By definition the ESA of a dipole is equal to

$$\sigma_1 = S_1 G_1, \tag{7.3}$$

where $S_1 = P_2/p$ – the ratio of the full power (P_2) reemitted by a dipole to the density of the power flux (p) of a plane wave incident on a dipole; G_1 – directive gain of a dipole.

For a dipole oriented at angle θ to the electrical vector $\bar{E}$ of an incident wave (Fig. 7.1), the reemitted power P_2 is equal to

$$P_2 = P_{20} \cos^2 \theta, \tag{7.4}$$

where P_{20} – the power emitted by a dipole in an equatorial plane (when $\theta = 0$).

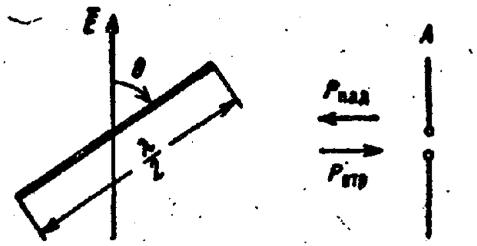

Fig. 7.1. A half-wave dipole randomly oriented in space.

As is known, the value of the power P_{20} can be found by the formula

$$P_{20} = \frac{1}{2} i^2 R_{\Sigma}, \tag{7.5}$$

where i – the amplitude of the current at the antinode; R_{Σ} – the resistance of dipole emission.

For a half-wave dipole

$$i = h_{\text{д}} \frac{E}{R_{\Sigma}}, \tag{7.6}$$

where E – the amplitude of an electrical field of a given plane wave; $R_{\Sigma} = 73.3\ \Omega$ – the resistance of a half-wave dipole emission; $h_{\text{д}} = \lambda/\pi$ – the effective height of a half-wave dipole.

From the above-cited relationships we find the power reemitted by a half-wave dipole

$$P_2 = \frac{1}{2} \frac{\lambda^2}{\pi^2} \frac{E^2}{73{,}3} \cos^2\theta \ [\mathrm{W}]. \tag{7.7}$$

The density of the power flux of an incident wave (the absolute value of an Umov-Poynting vector) is determined by the formula

$$p = \frac{E^2}{240\pi}. \tag{7.8}$$

Thus, from (7.3), (7.7) and (7.8) taking into account that, for a half-wave dipole $G = 1.65 \cos^2 \theta$ we finally obtain

$$\sigma_1 = 0{,}86\lambda^2 \cos^4 \theta. \tag{7.9}$$

With the coincidence of polarizations of a dipole and an incident wave the ESA of a half-wave dipole will be maximum

$$\sigma_{1\text{макс}} = 0{,}86\,\lambda^2. \tag{7.10}$$

In general θ is a random variable. With sufficient accuracy for practice taking into account the considerations expressed in the preceding paragraph, one may assume that the random variable θ is distributed with uniform density.

To determine the average value of the ESA of a dipole $\bar{\sigma}_1$ it is necessary to find the parameters of the law of distribution of the random variable θ.

The assumption of the equiprobable orientation of dipoles means that within the limits of any elementary solid angle $d\Omega_i$ (Fig. 7.2) the number of dipoles is approximately equal. An elementary solid angle $d\Omega_i$ can occupy any position with identical probability within the limits of the whole solid angle 4π. In a spherical system of coordinates the angle of elevation θ characterizes the position of each elementary solid angle $d\Omega_i$ and accordingly the distribution density is equal to

$$p(\Omega) = p(\theta) = \frac{1}{4\pi}. \tag{7.11}$$

Below we will consider that ESA of a cloud of dipoles does not depend on the relationship of the polarizations of the receiving and transmitting antennas, i.e., in other words, we will consider that their polarizations are identical. In general it is necessary to consider the difference of the polarizations.

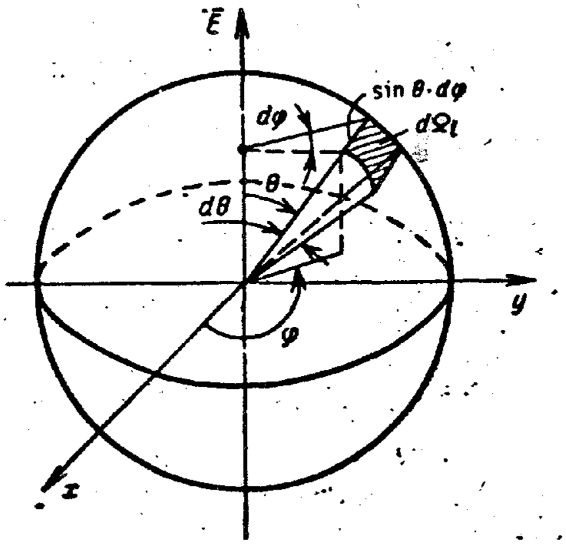

Fig. 7.2. Surface element in a spherical system of coordinates.

The probability of that a dipole will be within the limits of the elementary solid angle $d\Omega$ is equal to

$$p(\theta)\,d\Omega = \frac{d\Omega}{4\pi}. \tag{7.12}$$

In order to find the mean value of the ESA of a dipole (mathematical expectation of $\bar{\sigma}_1$), it is necessary to carryout averaging of the value of σ_1 determined by formula (7.9) in space for the whole solid angle $\Omega = 4\pi$

$$\bar{\sigma}_1 = \int_\Omega \sigma_1(\theta)\,p(\theta)\,d\Omega = \int_\Omega \sigma_1(\theta)\,\frac{d\Omega}{4\pi}. \tag{7.13}$$

In a spherical system of coordinates a surface element of a sphere of unit radius is equal to

$$ds = d\Omega = \sin\theta\,d\theta\,d\varphi. \tag{7.14}$$

Integrating (7.13), we will obtain

$$\bar{\sigma}_1 = \int_0^{2\pi} d\varphi \int_0^{\pi} \sigma_{1\text{макс}} \cos^4\theta\,\frac{1}{4\pi}\sin\theta\,d\theta. \tag{7.15}$$

Hence

$$\bar{\sigma}_1 = \frac{\sigma_{1\text{макс}}}{5} = 0{,}17\,\lambda^2. \tag{7.16}$$

Thus, the average ESA of a pack of dipoles $\bar{\sigma}_\text{П}$ will be equal to

$$\bar{\sigma}_\text{П} = N_\text{Пэ}\,\bar{\sigma}_1 = 0{,}17\lambda^2 N_\text{Пэ}, \tag{7.17}$$

where $N_\text{Пэ} = \eta N$ – the number of effectively acting dipoles in a pack.

Above we derived formula (7.16) for determining the ESA of a dipole, randomly oriented with respect to the direction of an electrical vector of an incident wave. In the process of deriving formula (7.16) it was assumed that the points of emission and reception coincide, and the angle θ between the electrical vector and the dipole is a random variable with an equiprobable law of distribution in space.

In a number of cases it is important to know the value of the ESA in a direction not coinciding with the direction to the source of the emission. Below the formula is derived for this case.

Let us assume that the direction of the reception of a signal reflected from a dipole forms angle ψ with the direction to the source of irradiation (Fig. 7.3). Let us designate by θ the angle between the dipole and the electrical vector of the irradiating field.[1]

Angle θ is a random variable with an equal law of distribution, i.e., the probability density p(θ) is determined by formula (7.11).

If $\sigma_{1\text{макс}} = 0{,}86\lambda^2$ – the ESA of a dipole when θ = 0, then for θ ≠ 0 and ψ ≠ 0

$$\sigma(\theta, \psi) = \sigma_{1\text{макс}} \cos^2\theta \cos^2(\theta + \psi). \tag{7.18}$$

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[1] It is assumed that the polarizations of the receiving and transmitting antennas coincide.

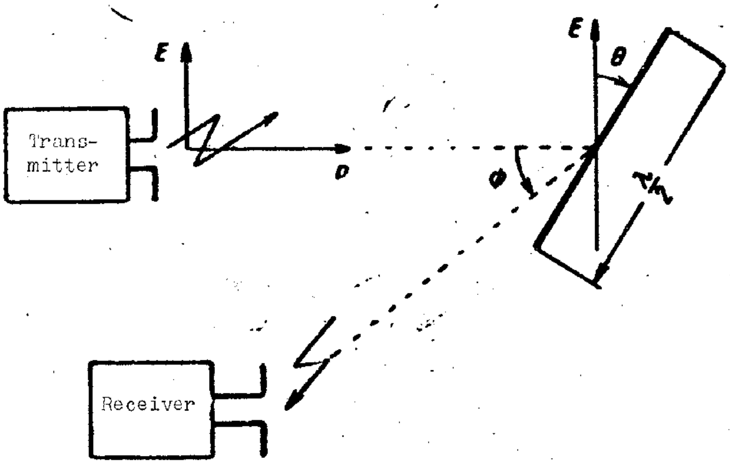


Fig. 7.3. A half-wave dipole randomly oriented in space with respect to the spaced receiver and transmitter.

It is interesting to us that the value of the ESA of one dipole is defined as a mathematical expectation $\overline{\sigma(\theta, \psi)}$, i.e.,

$$\sigma_\psi = \overline{\sigma(\theta, \psi)} = \int_\Omega \sigma(\theta, \psi) p(\theta) d\Omega. \tag{7.19}$$

The integration is carried out within the limits of the whole solid angle $\Omega = 4\pi$.

Replacing $\sigma(\theta, \psi)$ and $p(\theta)$ with their values from (7.18) and (7.11) and also considering the known expression for the differential of the solid angle of the unit radius $d\Omega = \sin\theta \, d\theta \, d\varphi$ (Fig. 7.2), we will obtain

$$\begin{aligned}\sigma_\psi &= \overline{\sigma(\theta, \psi)} = \int_\Omega \sigma(\theta, \psi) p(\theta) d\Omega = \\ &= \frac{\sigma_{1\text{макс}}}{4\pi} \int_0^{2\pi} \int_0^{\pi} \cos^2\theta \cos^2(\theta + \psi) \sin\theta \, d\varphi \, d\theta = \\ &= \frac{\sigma_{1\text{макс}}}{4\pi} \int_0^{2\pi} \int_0^{\pi} \cos^2\theta \cos^2(\theta + \psi) \sin\theta \, d\theta.\end{aligned} \tag{7.20}$$

The second integral of expression (7.20) by the transformation

$$\cos(\theta + \psi) = \cos\theta \cos\psi - \sin\theta \sin\psi$$

reduces to a sum of tabular integrals of the form:

$$I_1=\int_0^{\pi}\cos^4\theta\sin\theta\,d\theta=\frac{2}{5},$$

$$I_2=\int_0^{\pi}\cos^3\theta\sin^2\theta\,d\theta=0,$$

$$I_3=\int_0^{\pi}\cos^2\theta\sin^3\theta\,d\theta=\frac{4}{15}.$$

Thus, we will finally obtain

$$\sigma_\psi=\frac{\sigma_{1\text{макс}}}{5}\cos^2\psi+\frac{2}{15}\sigma_{1\text{макс}}\sin^2\psi \tag{7.21}$$

or

$$\sigma_\psi=0.17\lambda^2\cos^2\psi+0.11\lambda^2\sin^2\psi. \tag{7.22}$$

When $\psi = 0$ the well-known formula (7.16) is obtained.

Expression (7.22) makes it possible to draw the conclusion that the maximum dissipated power corresponds to the angles $\psi = 0$ and $\psi = \pi$, and the minimum – to the angles $\psi = \pi/2$ and $\psi = (3/2)\pi$. The value of σ_ψ in the direction of the minimum ($\psi = \pi/2$ or $(3/2)\pi$) corresponds approximately to $0{,}65\,\sigma_{1\text{макс}}$ (Fig. 7.4).

The obtained formulas are correct for an ideally conducting half-wave dipole. The real half-wave dipoles due to final conductivity and thickness possess greater ranginess than an ideal half-wave dipole. Increasing the length of a dipole reflector to values larger than half wave leads to a decrease in the ESA. However, when the values of dipole length are multiples of the number of half-waves, its ESA again increases and can be somewhat larger than for a half-wave dipole (Fig. 7.5). The latter does not mean that dipole reflectors intended for the suppression of longer-wave stations will also be effective against radar operating on shorter waves. The fact is that the number of dipoles in packs for longer waves is decreased, and accordingly the ESA of a pack for shorter waves is decreased.

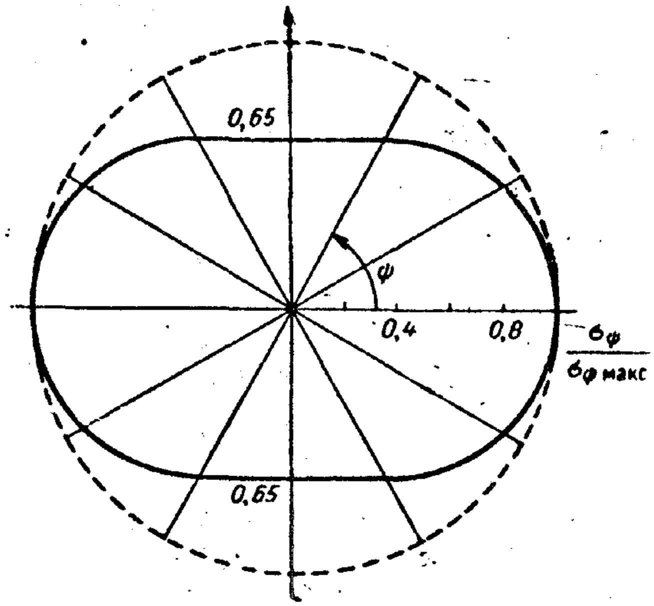


Fig. 7.4. The dependence of the ESA of a half-wave dipole on the angle between the direction to the transmitter and receiver.

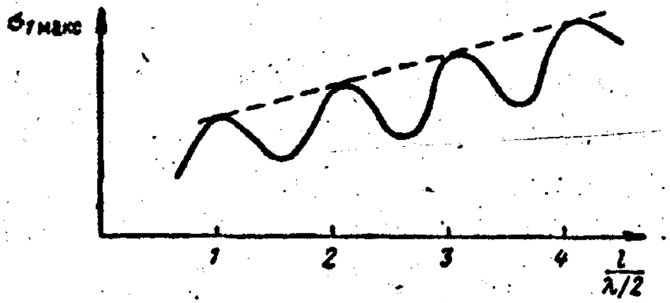


Fig. 7.5. The dependence of the ESA of a dipole on its relative length.

The qualitative dependence of the ESA of a dipole on its relative length is shown in Fig. 7.5, where along the axis of the abscissas there is plotted the ratio of the length of the dipole to half of the wavelength.

7.4. The Fluctuation Spectrum of Signals Reflected from Dipoles

Atmospheric turbulence, the aerodynamics of dipoles and other factors affect the rate of scattering of a cloud of dipoles in space.

Due to this the amplitude of the resultant reflected signal varies in time according to a random law, furthermore, an expansion of the frequency spectrum of the total reflected signal occurs. The expansion of the spectrum of the total reflected signal occurs due to the Doppler components of the spectrum as a result of the following causes:

- inherent scattering of the dipole reflectors in the atmosphere;
- scattering in space due to atmospheric turbulence;
- movement due to wind;
- the descent of the dipoles due to the influence of gravity;
- scattering in space due to the turbulent stream of the aircraft.

Furthermore, the causes of the broadening of the fluctuation spectrum of the amplitude of the reflected signal are: the characteristic rotation of the dipoles, the irregularity of the radiation pattern of the radar antenna and others.

The standardized function of the spectral density of the fluctuations of amplitudes of the reflected signal due to the shifts of the reflectors in accordance with the investigations of A. A. Zagorodnikov [62] equalled

$$G(F) = \exp\left(-\frac{\lambda^2 F^2}{16\overline{v_x^2}}\right), \tag{7.23}$$

where λ – the wavelength of the suppressed radar; F – frequency;

$$\overline{v_x^2} = \overline{(v_x^2)_\text{а}} + \overline{(v_x^2)_\text{б}} + \overline{(v_x^2)_\text{в}}; \tag{7.24}$$

v_x – the component of speed of the reflector along the radar Ox axis (Fig. 7.6); $(v_x)_\text{а}$ – the component of the rate of descent of the dipoles due to their own weight; $(v_x)_\text{б}$ – the component of speed of shift of the dipoles due to the wind; $(v_x)_\text{в}$ – the component of speed of the dipoles due to their shift under the effect of atmospheric turbulence.

From (7.23) we find the spectrum width at a level of 0.5 (on the output of amplitude detector)

$$F_{0,5}=\frac{3,33}{\lambda}\sqrt{\overline{v_x^2}}=\sqrt{F_a^2+F_{\text{б}}^2+F_{\text{в}}^2},$$

where

$$F_a^2=\frac{11,5}{\lambda^2}\overline{(v_x^2)_a};$$

$$F_{\text{б}}^2=\frac{11,5}{\lambda^2}\overline{(v_x^2)_{\text{б}}};$$

$$F_{\text{в}}^2=\frac{11,5}{\lambda^2}\overline{(v_x^2)_{\text{в}}}.$$

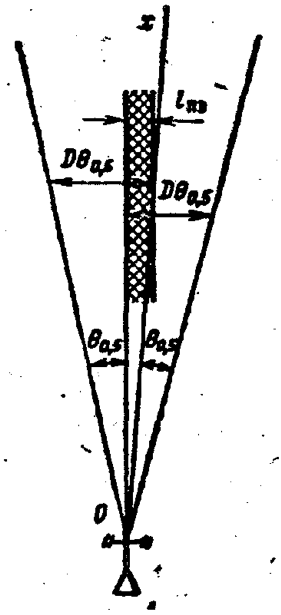


Fig. 7.6. A variant of the distribution of a band of dipole reflectors with respect to the radar beam: $D\theta_{0.5}$ – the linear resolving power of the radar by angle; $l_{\text{нэ}}$ – effective bandwidth of the dipole reflectors.

The broadening of the F_a spectrum due to the diversity of the rates of descent is caused by the presence of dipoles descending at different speeds in the cloud. As an illustration there is depicted in Fig. 7.7 the distribution function of the rates for descending reflectors. The values shown on the ordinate axes indicate the relative number of dipoles in the cloud, having a rate of descent v. The figure shows that there are two stable groups of dipoles – the "slow" and the "fast." The presence of the "slow" dipoles is explained by the fact that these dipoles tend to mainly assume a horizontal orientation. A portion of the dipoles due to shallow indentations, deformations, etc., has a similarity to aerodynamic elevators, which ensure their stable descent chiefly in a vertical orientation. This portion of the dipoles forms the "fast" group.

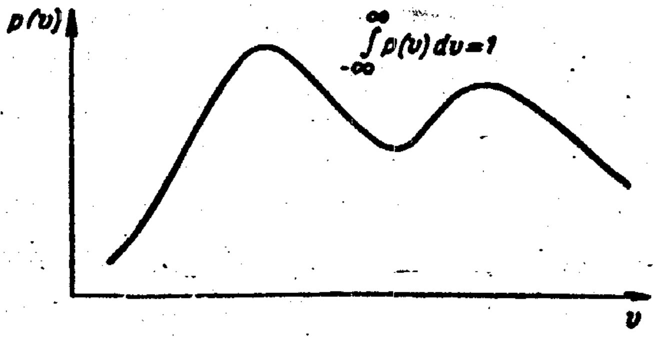


Fig. 7.7. Velocity distribution density of descending dipoles.

The character of the movement of the "slow" and "fast" dipoles is shown in Fig. 7.8. Experimental data attest to their predominant horizontal distribution [58].

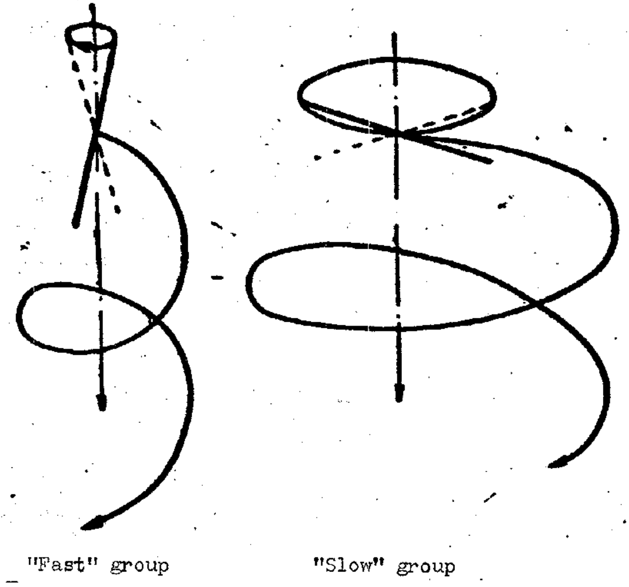


Fig. 7.8. The travel pattern of the "slow" and "fast" dipoles.

The component of the F_a spectrum due to the diversity of the rates of descent is maximum for a cloud with small dimensions as compared to the dimensions of the pulse volume of suppressed radar. When the dimensions of the cloud are increased this component gradually decreases to a certain constant value, the magnitude of which is greater the greater the turbulence of the atmosphere. The value of the component of F_a essentially depends on the angle of elevation of the cloud. For radar operating at small angles of elevation, it cannot be considered.

The component of the $F_б$ spectrum due to the shift of the dipoles by the effect of the wind is proportional to the diversity of windspeeds along the vertical (gradient variation of windspeed) and increases as the dimensions of the cloud increase. The average windspeed transporting the cloud as a whole causes a shift in the spectrum and only an insignificant broadening of it. For small clouds the $F_б$ component increases directly proportional to the dimensions of the cloud. With considerable cloud dimensions exceeding the dimensions of the antenna radiation pattern with respect to the angle of elevation, the width of the $F_б$ spectrum does not depend on the cloud dimensions. This component can attain a significant value with the creation of radar interferences, if the clouds of dipoles have large vertical dimensions.

The component due to atmospheric turbulence practically does not depend neither on the dimensions nor on the angular coordinates of the cloud, and is determined mainly by the meteorological parameters of the atmosphere. It can be approximately considered

$$F_в = \frac{v_в}{4\lambda},$$

where $v_в$ – the averaged windspeed; λ – wavelength of the suppressed radar.

The spectrum width of the reflected signals is significantly affected by these meteorological parameters of the atmosphere:

- average windspeed;
- vertical gradient of windspeed;
- average temperature of the air layer;
- vertical gradient of temperatures and atmospheric turbulence.

The indicated meteorological parameters of the atmosphere vary with height. Consequently, the spectrum width of reflected signals at a given distribution of meteorological parameters depends on height. In Fig. 7.9 there is depicted the dependence of the standardized spectral density of the fluctuations of signals reflected by a cloud of dipoles on the product of Fλ (F – the frequency of fluctuation in Hz, λ – the wavelength in centimeters [61]). The dependence was experimentally plotted for certain meteorological conditions.

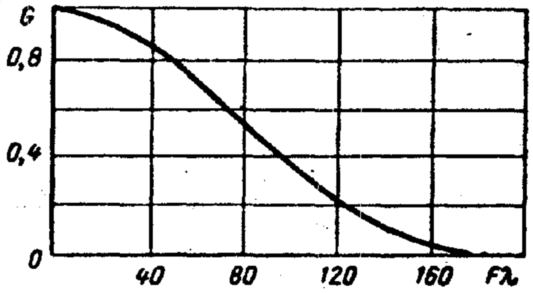


Fig. 7.9. The standardized spectral density of an envelope of signals reflected by a cloud of dipoles.

7.5. Weakening of Radio Waves in Clouds of Dipoles with Large Concentrations

It is known that electromagnetic waves passing through a cloud of dipoles experience weakening due to the scattering of energy by the dipoles. Let us determine the coefficient of absorption of radio waves in a cloud of dipole reflectors with a concentration of $\bar{n}$ dipoles per unit of volume (m^3).

In the first approximation it is possible to assume that the elementary volume of a cloud with an area of 1 m^2 and a thickness of dx (Fig. 7.10) disperses energy proportional to its ESA, i.e.,

$$dP = -P\bar{\sigma}_{э0}dx, \tag{7.25}$$

where dP – the power dispersed by the elementary volume; P – the power of the signal at the "input" of the elementary volume; $\bar{\sigma}_{э0}$ –

the specific ESA $\left[\frac{м^2}{м^3}\right]$ (ESA of the dipoles distributed in a unit of volume).

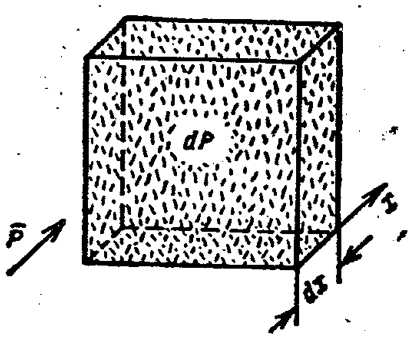


Fig. 7.10. Elementary volume of a cloud of dipoles.

Taking into account that the specific ESA is equal to

$$\bar{\sigma}_{э\,0} = \bar{n}\,0{,}17\lambda^2, \tag{7.26}$$

equation (7.25) can be written in the form

$$\frac{dP}{dx} + P\bar{n}\,0{,}17\lambda^2 = 0. \tag{7.27}$$

Integrating equation (7.27) considering the boundary conditions ($P = P_0$ when $x = 0$) we find the power lost in the passage of the electromagnetic waves through a cloud with a thickness of x meters:

$$P = P_0 e^{-\bar{n}\,0{,}17\lambda^2 x}. \tag{7.28}$$

Here P_0 — a certain initial power (the power of the "input" of cloud).

From expression (7.28) we can easily find the attenuation factor β:

$$\beta = 0{,}73\lambda^2\bar{n} \text{ [dB/m]},$$

where λ — wavelength in meters.

The given relationship for the attenuation factor β makes it possible to write the dependence of power on distance in the following form:

$P = P_0 10^{-0,1\beta x}$ – for the case, when the radio waves pass through the cloud only in one direction;

$P = P_0 10^{-0,2\beta x}$ – for the case, when the radio waves pass through the cloud twice (out and back).

Let us give an example of the calculations.

We will consider that a cloud with a thickness of x_0 is shielding, if the transmission range of the radar decreases by 10 times. Going on these assumptions the absorption should be 40 dB. Let us determine the necessary cloud concentration and dimension with respect to a specific radar (λ = 0.03 m).

If the dimensions of the cloud are limited in length to one kilometer ($x = 10^3$ m), then the total attenuation of 40 dB will take place at a specific attenuation factor β = 0.02 dB/m. Hence

$$\bar{n} = \frac{0,02}{0,73 \cdot 9 \cdot 10^{-4}} \approx 30 \text{ dyn/m}^3.$$

This concentration of dipoles is very high.

7.6. The Peculiarities of the Suppression of Detection and Target Designation Pulsed Radar by Passive Interferences

In appraising the effect of passive interferences created with dipole reflectors on pulse radar, it is important to know the ESA generated by the dipole reflectors located in the pulsed volume. For this it is necessary to calculate the number of dipoles getting into the pulsed volume of the radar (Fig. 7.11). The borders of pulsed volume V are determined by the pulse width and the beam width of the antenna of the suppressed radar, i.e.,

$$V \approx D\theta_{0,5} D\Phi_{0,5} \frac{c\tau}{2},$$

where D – the distance of the pulsed volume from the radar;

$\Phi_{0,5}$, $\theta_{0,5}$ – the width of the antenna radiation pattern of the suppressed radar by azimuth and angle of elevation respectively at a level of half power; τ – the pulse width of the suppressed radar; c – the speed of the electromagnetic waves in free space.

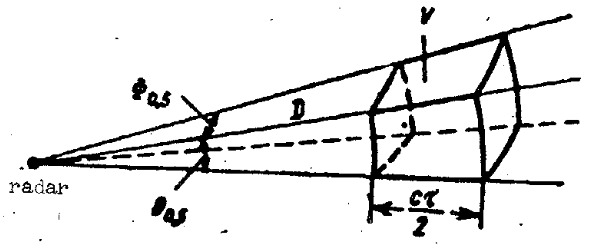


Fig. 7.11. Pulsed volume of radar.

If it is considered that the dipoles are distributed in the cloud uniformly with an average volume density $\bar{n}$, then the ESA of the dipoles located in the pulsed volume can be determined by the formula

$$\bar{\sigma}_{\text{BO}} = \bar{n}V\bar{\sigma}_1 = \bar{n}D^2\theta_{0,5}\Phi_{0,5}\frac{c\tau}{2}\bar{\sigma}_1. \tag{7.29}$$

Expression (7.29) shows that the effectiveness of passive interferences depends essentially on the resolving power of the suppressed radar for angles and distance.

Above we considered that the dipoles are distributed in the cloud uniformly on the average. However in practice after the opening of a pack the dipole density in a cloud varies in time in accordance with the laws of turbulent diffusion. The dipoles experience random flight, mainly, due to atmospheric turbulence. The dimensions of the cloud with time increase, and the average density of the dipoles in it decreases.

In Fig. 7.12 there is presented the dependence of the density of dipole distribution at the point of the opening of a pack on one coordinate x. This figure gives a qualitative picture of the change of concentration of the dipoles in time. Sufficiently exact quantitative characteristics of the volume density distribution of dipoles

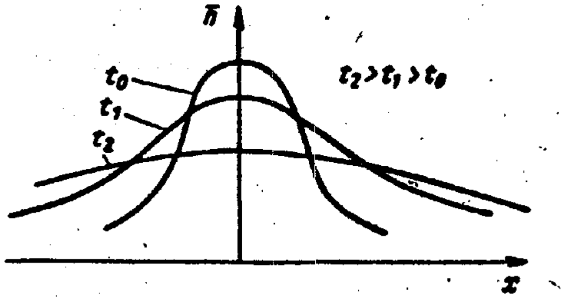


Fig. 7.12. The density of dipole distribution along one coordinate.

can be obtained by solving the equation of turbulent diffusion.

Depending on the value of dispersion $\sigma_д^2$ of dipole distribution in a turbulent medium the coefficient of turbulent diffusion D varies differently in time, where

$$D=\frac{1}{2}\frac{\partial}{\partial t}\sigma_д^2.$$

As the investigations of A. A. Zagorodnikov [62] showed, the distribution of dipole reflectors in a turbulent atmosphere with an effective value of cloud width $\sigma_д \leqslant 25$ м is determined by the coefficient of turbulent diffusion, time-dependent on the quadratic law (the Kolmogorov-Obukhov law of turbulent diffusion):

$$D=\frac{1}{2}\frac{\partial}{\partial t}\sigma_д^2=\frac{1}{2}\frac{\partial}{\partial t}\left(\eta t^{\frac{3}{2}}\right)^2=\eta_1 t^2,$$

$$\sigma_д=\eta t^{\frac{3}{2}},$$

where η, η_1 – proportionality factors.

If however 25 м $\leqslant \sigma_д \leqslant 500$ м, then the coefficient of turbulent diffusion is $D \backsim t$ and $\sigma_д \backsim t$; accordingly with large values of effective cloud width ($\sigma_д > 500$ м) normal diffusion occurs ($D=\mathrm{const}$, $\sigma_д \backsim \sqrt{t}$).

In practice the second case is of the greatest interest $(25\ \text{м} \leqslant \sigma_{\text{Д}} \leqslant 500\ \text{м})$. For the indicated conditions the solution of the equation of turbulent diffusion reduces to the following formula for the distribution density:

$$p(V)\,dV = \frac{1}{\sqrt{2\pi}\,\sigma_{\text{Д}\,ox}t}\exp\left[-\frac{(x-x_0)^2}{2\sigma^2_{\text{Д}\,ox}t^2}\right]\times$$
$$\times\frac{1}{\sqrt{2\pi}\,\sigma_{\text{Д}\,oy}t}\exp\left[-\frac{(y-y_0)^2}{2\sigma^2_{\text{Д}\,oy}t^2}\right]\frac{1}{\sqrt{2\pi}\,\sigma_{\text{Д}\,oz}t}\times$$
$$\times\exp\left[-\frac{(z-z_0)^2}{2\sigma^2_{\text{Д}\,oz}t^2}\right]dV. \tag{7.30}$$

Here $p(V)dV$ – the probability that after time t a given dipole will appear in the elementary volume dV;

$$x_0 = C_x t;$$
$$y_0 = C_y t;$$
$$z_0 = C_z t;$$

C_x, C_y, C_z – the speed (speed of flow) of air respectively on axes x, y, z;

$$\sigma_{\text{Д}\,ox} = 0{,}0168C'_x;$$
$$\sigma_{\text{Д}\,oy} = 0{,}0168C'_y;$$
$$\sigma_{\text{Д}\,oz} = 0{,}0168C'_z;$$

C'_x, C'_y, C'_z – the averaged windspeed respectively on axes x, y, z.

It is possible to arrive at this formula by direct determination (on the basis of considering the random flights of a particle) of the probability that a particle (dipole) after a certain time will appear within limits of the elementary volume dV.

Knowing the law of dipole distribution in space p(V), it is possible to determine the number of dipoles in pulsed volume V

$$N_{\text{ио}} = \iiint\limits_V N_{\text{п э}} n_{\text{п}} p(V)\,dV, \tag{7.31}$$

where $n_{\text{п}}$ — the number of simultaneously ejected packs; $N_{\text{пэ}}$ — number of effectively acting dipoles in a pack.

The ESA of dipoles located in a pulsed volume (ESA of the pulsed volume), accordingly is found by the formula

$$\bar{\sigma}_{\text{и о}} = N_{\text{ио}} \bar{\sigma}_1. \tag{7.32}$$

A more exact calculation of the ESA of the pulsed volume requires a calculation of the form of the antenna radiation pattern in both planes (along θ and Φ), and also the form of the sounding pulse $\tau(t)$. The number of dipoles in pulsed volume $N_{\text{ио}}$ in this case is determined by the triple integration along θ, Φ and τ of the distribution density $p(V)$ multiplied by the appropriate weight functions $F(\theta, \Phi)$ and $\tau(t)$, considering the real form of the pulsed volume.

The target is not detected in a cloud of windows [dipole interferences], if the power of the interference signals (reflected from the dipoles distributed in the pulsed volume) exceeds by a definite number of times the power of the effective signal (reflected from the target). The ratio of the power of the interference signal to the effective signal at the input of the receiver equalled

$$k = \left(\frac{P_{\text{п}}}{P_{\text{c}}}\right)_{\text{вх}} = \frac{\bar{\sigma}_{\text{ио}}}{\bar{\sigma}_{\text{ц}}}, \tag{7.33}$$

where $\bar{\sigma}_{\text{ио}}$ — the ESA of the pulsed volume; $\bar{\sigma}_{\text{ц}}$ — the ESA of the target.

The problem of detecting the effective signal against a background of passive interferences has much in common with the problem of detecting a signal in Gaussian noise. Signals, reflected from a cloud of dipoles, when it has a sufficiently large density, due to the central limiting theorem can be examined as Gaussian noise; however in contrast to white noise the autocorrelation function of this noise will not coincide with the δ-function. This circumstance has an effect on the method of calculating the probability ratio

$$\Lambda_1(u) = \frac{p_1(u)}{p_0(u)},$$

where $p_0(u)$ and $p_1(u)$ – the distribution density respectively of noise and additive mixture of signal and noise.

The methodology of determining the threshold ratio, corresponding to the optimum solution, naturally, remains the same, as in detecting a signal against a background of active noise interferences.

Let us note that due to the correlation of noise generated by the reflections from the cloud of dipoles, the coefficient of suppression by passive interferences will depend on the parameters of the autocorrelation function or what is the same thing, on the parameters of the spectral density of signals reflected from the cloud of dipoles. This is especially important to consider in determining the coefficient of suppression of the radar, having attachments for compensating signals from passive interferences.

The minimal necessary ratio $\left(\frac{P_\text{п}}{P_\text{c}}\right)_\text{мин}$, where the probability of correct detection of a target against a background of dipole reflectors for a given probability of false alarm is less than a certain value (0.1-0.5), is called the coefficient of suppression of pulsed radar by passive interferences.

The determination of the ESA of a pulsed volume with (7.31) and (7.32) – a rather unwieldy problem, requiring great calculating operations. In practice sometimes it is expedient to use simpler methods for calculating the ESA of a reflecting pulsed volume.

Simplified Methods of Calculating the ESA of a Pulsed Volume

An important characteristic of a cloud of dipole reflectors is its effective width $l_\text{пэ}$. The effective width of a cloud is determined by the region, within whose limits 70% of all ejected dipoles are contained.

If one considers the law of distribution of dipoles along any coordinate for a given moment of time normal, then the effective width of the cloud $l_{\text{пэ}}$ is equal to (Fig. 7.13)

$$l_{\text{пэ}} = 2\sigma_{\text{д}},$$

where $\sigma^2_{\text{д}}$ – the dispersion of the distribution density p(x) of the dipoles on a given coordinate x.

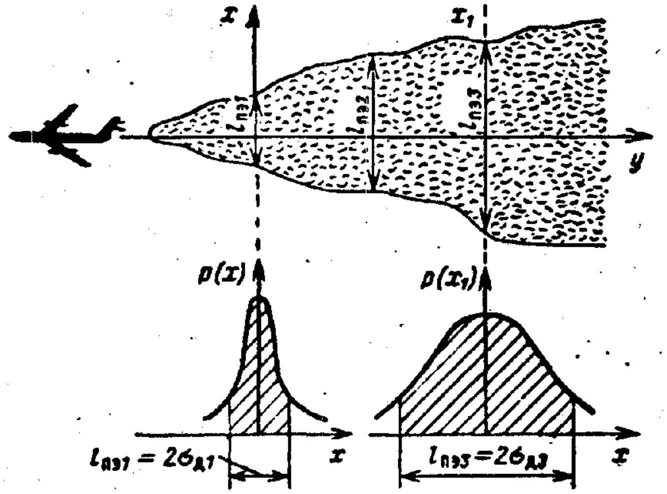


Fig. 7.13. The process of the development of a dipole cloud in space.

As a rule, in practice the effective width of a cloud frequently does not exceed all or part of the dimensions of a pulsed volume of suppressed radar.

Most frequently

$$l_{\text{пэ}} < D\theta_{\text{о.з}} \text{ and } D\Phi_{\text{о.з}},$$

where $D\Phi_{\text{о.з}}$, $D\theta_{\text{о.з}}$ – the linear resolving power of the suppressed radar respectively for aximuth and angle of elevation.

Let us examine a case, when a covered aircraft PS_1 flies in a band of dipole reflectors in the direction of a suppressed radar (Fig. 7.14a). Flight in the indicated direction is frequently unfavorable from the point of view of jamming, since the value of the ESA of the pulsed volume σ_{01} can be the least when compared to all the

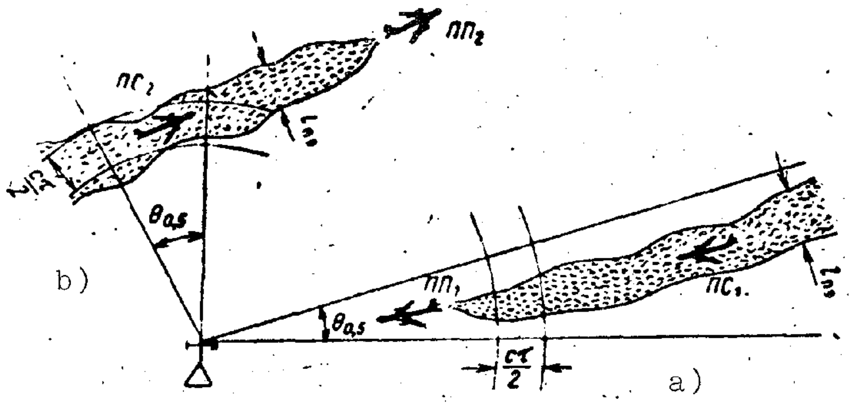
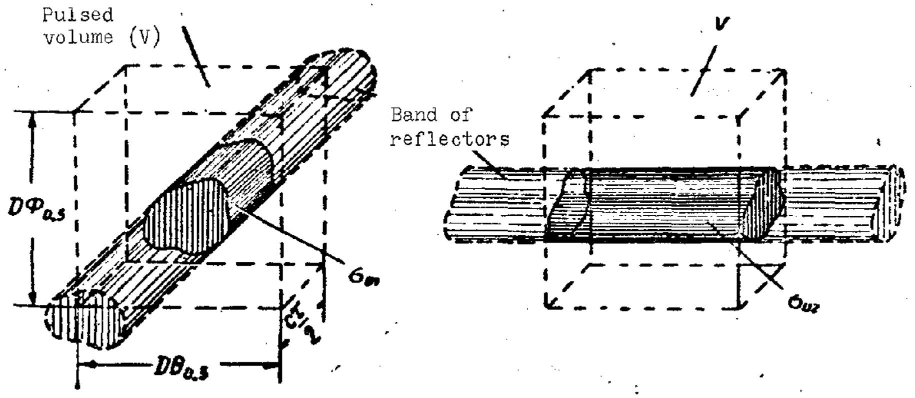


Fig. 7.14. The effect on radar of interferences created by bands of dipole reflectors: a) "longitudinal" band of dipoles; b) "transverse" band of dipoles.

other directions of flight. For example, in the flight of the aircraft (PS_2 and PP_2) at a certain course angle [heading] to the radar the value of the ESA of the pulsed volume σ_{02} will be greater as compared to σ_{01} (Fig. 7.14 and 7.15). Consequently, the safeguarding of the conditions of camouflage will also be an easier matter.[1]



a) b)

Fig. 7.15. The number dipoles located in a pulsed volume of radar for the "longitudinal" (a) and the "transverse" (b) bands.

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[1] In this case the variation of the ESA of the aircraft depending on its altitude with respect to the suppressed radar is not considered.

Let us assume that packs of dipole reflectors are uniformly dropped by an interference producer (PP) at a rate of $t_{\text{п}}=\text{const}$. The jammer [PP] flies at a constant speed $v_{\text{п}}$. If each time the jammer simultaneously drops $n_{\text{п}}$ packs, then per unit of path after the establishment of the diffusion process (the coefficient of diffusion increases approximately proportional to time or is constant) the number of dipoles on the average will be

$$N=\frac{n_{\text{п}}N_{\text{пэ}}}{v_{\text{п}}t_{\text{п}}},$$

where $N_{\text{пэ}}$ – the number effectively operational dipoles in a pack.

If the condition $D\theta_{0,5};\ D\Phi_{0,5}>l_{\text{пэ}}$ is fulfilled and the reflectors along the bands are distributed evenly on the average, then the average number of dipoles in a pulsed volume will be defined as the product of the pulse width expressed in units of length $\left(\frac{c\tau}{2}\right)$ per average specific density of dipoles in a band

$$N_{\text{и о}}=\frac{c\tau}{2}\ N=\frac{c\tau}{2}\frac{n_{\text{п}}N_{\text{пэ}}}{v_{\text{п}}t_{\text{п}}}.$$

Accordingly the ESA of a pulsed volume equalled

$$\bar{\sigma}_{\text{и о}}=\bar{\sigma}_1 N_{\text{и о}}=0{,}17\lambda^2\frac{c\tau}{2}n_{\text{п}}\frac{N_{\text{пэ}}}{v_{\text{п}}t_{\text{п}}}. \qquad (7.34)$$

It is necessary to note, that formula (7.34) is correct under the condition of the uniform ejection of the dipoles from the aircraft – the jammer [PP] and with the bearing of the aircraft approximately in the direction of the suppressed radar (the suppressed radar, the band of dipole reflectors and the covered aircraft are in alignment). In the general care of a random position of the covered aircraft and of the cloud the ESA of the reflecting pulsed volume can be determined by formulas (7.31) and (7.32).

Knowing the coefficient of suppression $k_{\text{п}}$ of a specific pulsed radar with passive interferences, it is possible to determine the number of dipole reflectors necessary to camouflage the covered aircraft.

The ESA of the pulsed volume necessary to camouflage the aircraft is equal to

$$\bar{\sigma}_{\text{п о}} = k_{\text{п}} \bar{\sigma}_{\text{ц}}, \tag{7.35}$$

where $\bar{\sigma}_{\text{ц}}$ – the ESA of the covered aircraft.

From (7.34) taking into account (7.35) we find the number of packs of dipole reflectors necessary to cover an aircraft for an interval of length L, assuming that their ejection occurs place in each pulsed volume,

$$n_{\text{п}} = \frac{\bar{\sigma}_{\text{ц}} k_{\text{п}} v_{\text{г}} t_{\text{п}}}{0.17\lambda^2 \frac{c\tau}{2} N_{\text{п э}}} \frac{L}{\frac{c\tau}{2}} = \frac{23.5 \sigma_{\text{ц}} k_{\text{п}} v_{\text{г}} t_{\text{п}} L}{\lambda^2 c^2 \tau^2 N_{\text{п э}}}. \tag{7.36}$$

Let us assume, for example, that it is necessary to cover with passive interference for a section with a length L = 100 km an aircraft, having an ESA of $\sigma_{\text{ц}} = 50\ м^2$.

We will consider: $k_{\text{п}} = 2$, $v_{\text{п}} = 200$ m/s, $t = \frac{c\tau}{2v} = 0{,}75$ s, $\lambda = 10\ см$, $\tau = 10^{-6}$ s.

We also assume that the dipoles have a pack ESA equal to

$$\sigma_{\text{п}} = 0{,}17\lambda^2 N_{\text{п э}} = 50\ м^2.$$

Under these conditions from (7.36) we obtain $n_{\text{п}} \approx 1330$ packs. Knowing the weight of each pack, it is easily possible to calculate the weight of dipoles necessary to camouflage an aircraft on an indicated flight route.

Effective Width of a Camouflaged Region

The section of space excluded from observation is called the camouflaged region. By the effective width of a camouflaged region $L_{\text{з}}$ is understood the mathematical expectation of the length of the interval between two boundary positions of the target with respect to the dipole cloud, at each point of which there is provided a necessary value of the interference signal ratio, with respect to the conditions of suppression. The radar cannot detect an aircraft located within the limits of the camouflaged region. The important characteristics of the camouflaged region are its dimensions, since they also strictly determine the information damage inflicted by the passive interferences. The effective width of the camouflaged region $L_э$ is determined not only by the width of the band of dipole reflectors $l_{пэ}$, but also by resolving power of the suppressed radar with respect to range and angle. The effective width $L_э$ depends also on the relative location of the band of dipoles and the suppressed radar. Thus, if the band of dipole reflectors is in the range of the main lobe of the antenna radiation pattern of the suppressed radar (longitudinal band), then the effective width of the camouflaged region can be approximately determined by the relationship (Fig. 7.6)

$$L_э \approx D\theta_{0,5} + l_{пэ}, \tag{7.37}$$

where $D\theta_{0,5}$ – the linear resolving ability of the suppressed radar with respect to angle; $l_{пэ}$ – the effective width of the band of dipole reflectors.

For example, with a beam width of $\theta_{0,5}=2^\circ$ at a distance of D = = 100 km a band of dipole reflectors with $l_{пэ}=300$ м creates a camouflaged region with a width $L_э=3{,}8$ км.

From the conclusion of formula (7.37) an assumption was made about necessity of all the dipoles situated in a given cross section of the band getting into the pulsed volume of the radar, when the maximum beam of the antenna was directed towards a conditional boundary of the band. This means that the width of a camouflaged region $L_э$ can be decreased by increasing the effective width of the band $l_{пэ}$. If however the specific density of the dipoles in the band is sufficiently large and it is not necessary that all the dipoles located in this section get into the pulsed volume in a given cross section, then an approximate relationship will occur between L_a and $l_{п э}$.

Thus, the effective width of the camouflaged region does not coincide with the effective width of the band of dipole reflectors. It increases with the increase of distance D.

Formula (7.37) is approximate. It does not consider a change in the interference ratio at the input of the suppressed radar depending on the form of the antenna radiation pattern. The calculation of the effect of the form of the antenna radiation pattern on the effective width of the camouflaged region was performed by V. T. Borovik.

In Fig. 7.16 there is shown the variation in power of the interference and effective signals with various angular positioning of the antenna with respect to the cloud of interferences and the target covered by it. The effective width of the camouflaged region is determined by the level of the power of the interference signal, the relationship of which to the power of the effective signal taken as the maximum of the antenna radiation pattern, was equal to the coefficient of suppression.

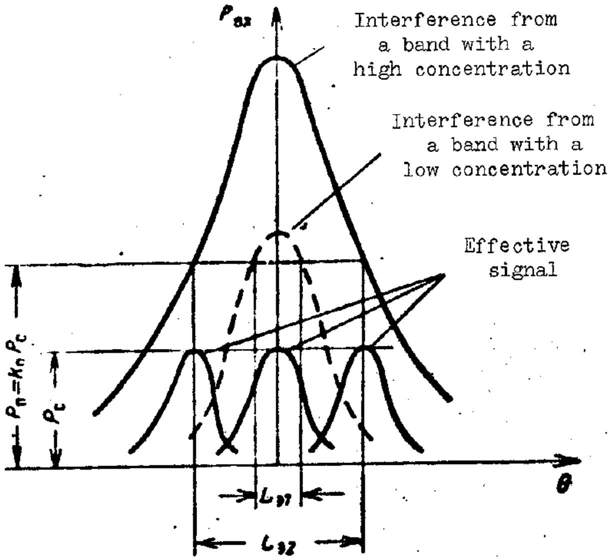
$P_{вх}$
Interference from a band with a high concentration
Interference from a band with a low concentration
Effective signal
$P_п = k_п P_c$
P_c
θ
$L_{э1}$
$L_{э2}$

Fig. 7.16. Variation of the power of the interference and effective signals with various angular positioning of the antennas relative to the cloud of interferences and the target covered by it.

7.7. Peculiarities of the Effect of Passive Interferences on Pulse-Coherent Radar

The protection of pulse radar with a large off-duty factor from passive interferences at present is, mainly, provided by alternate-period compensation of reflections from a cloud of dipoles [60].

The simplest type of block diagram of pulse radar with alternate-period compensation is shown in Fig. 7.17. Here M is a modulator forming short pulses for modulation of the high-frequency generator Г and long pulses for triggering the coherent oscillator [CO] for a time, approximately equal to the pulse repetition period T. Usually this time is somewhat shorter, in order to make it possible to damp the inherent oscillations in the circuits of the coherent heterodyne toward the beginning of the next cycle of operation. The coherent heterodyne is synchronized in phase with the high-frequency pulse of the generator, thereby the coherence of the signal of the heterodyne and radiated signal is ensured for a time, approximately equal to T.

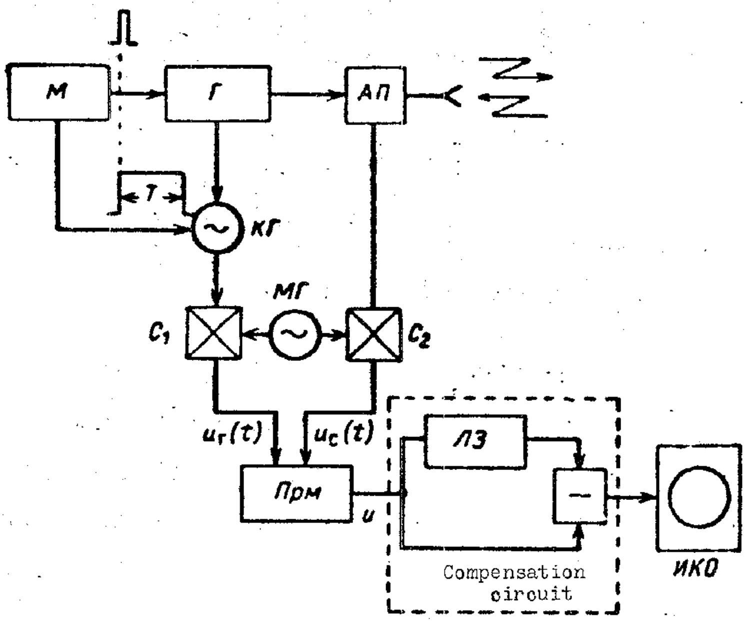

Fig. 7.17. Block diagram of a pulse-coherent radar.

The following, new, high-frequency pulse again carried out (for itself) the phasing of the coherent heterodyne. A similar kind of radar with a large off-duty factor is called pulse-coherent radar, and sometimes due to the short time of coherence it is called pseudo- or quasi-coherent radar.

The reflected signal and the signal of the coherent heterodyne after conversion in the mixers C_1 and C_2 enter into the input of the receiving device Prm (i-f amplifier, detector, video amplifier).

In Fig. 7.18 there are shown the sequence of the video pulses on the input of the compensation circuit for two cases: a nonmoving object and a moving target.

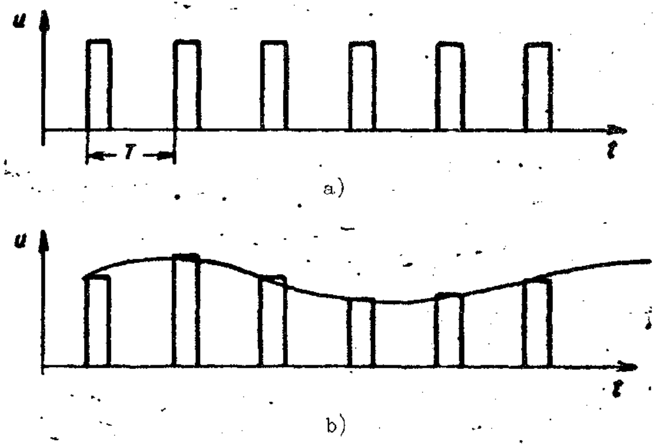

Fig. 7.18. Sequences of video pulses on the input of the compensation circuit for nonmoving (a) and moving (b) targets.

From the output of the receiver the video pulses enter the compensation circuit, consisting of a delay line LZ with a delay time T and a difference circuit. The resultant signal on the output of an ideally operating quasi-coherent radar will be equal to zero, if the target is stationary, and different from zero, if it is not moving with "blind" speed.[1]

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[1] Blind speeds mean target speeds, at which for a time equal to the period of the pulse sequence, the target traverses in a radial direction relative to the given radar a distance, which is a multiple of the whole number of half-waves. Due to this the phase lead during the time between the adjacent high-frequency pulses entering the input of the radar is an even multiple of π. This, finally, leads to the compensation of the effective signals.

In practice a cloud of dipoles cannot be considered absolutely stationary. Due to turbulent diffusion and other causes the individual dipoles of the cloud shift relative to each other in a random manner. Furthermore, due to the effect of the wind the cloud moves progressively as a single whole. The progressive movement of the cloud is usually taken into account in a quasi-coherent radar with the help of a so-called wind compensation circuit. (In the simplest case this circuit ensures a linear (in time) variation of the phase of high-frequency oscillations). Subsequently we shall talk about the influence of only the random shifts of the dipoles in the cloud. In monochromatic sounding signal after reflection from a cloud of dipoles moving in a random manner loses its monochromaticity and essentially becomes quasi-harmonic oscillation

$$u = u_{\text{п}}(t)\cos[\omega_0 t + \varphi_{\text{п}}(t)],$$

where $u_{\text{п}}(t)$ and $\varphi_{\text{п}}(t)$ – slowly varying as compared to $\omega_0 t$ are random functions of time.

The quality of the suppression by the alternate-period compensation circuit of the signals reflected from the cloud of dipoles, is mainly affected by the variations of the phase of the interference signal, occurring for a time equal to the pulse repetition period. The variations of the phase of the interference signal result in the corresponding video pulse of the resultant oscillation (of the interference signal and coherent oscillator) being different in amplitude from the adjacent pulse. For this reason there will not take place complete compensation of the signals reflected from the cloud of dipoles.

Increasing the multiplicity of subtraction (the use of two-three-fold, etc., subtraction) somewhat increases the effectiveness of the compensation circuit, however this does not completely solve the problem. Furthermore, the quality of the suppression of the signals reflected from a cloud of dipoles is affected by the instabilities of the transmitter frequency, the local oscillator, the coherent oscillator, the operational instabilities of the subassemblies, and also the fluctuating interferences.

The presence of uncompensated remainders makes it possible by increasing the density of the dipoles to overcome the effect of the alternate-period difference circuit and to provide camouflage for the moving targets with a cloud of passive interferences.

Let us estimate the value of the coefficient of suppression by the passive jamming of radar with alternate-period compensation. With respect to this determination by the coefficient of suppression $k_{\text{п}}$ is understood the minimum necessary ratio of interference power to signal power on the receiver input (in the passband of the linear part), at which the probability of correct detection does not exceed a given value for a given probability of false alarm.

Let us assume that for radar without moving-target selection SDTs circuits such a ratio is known and let us assume that with the ratio $\frac{P_{\text{п}}}{P_{\text{с}}}=k_{\text{п}}$ on the input the interference ratio on the output of the receiving device, i.e., on the input of the indicator (or other terminal device), was equal to $k_{\text{в}}$. When $P_{\text{п}}$ and $P_{\text{с}}$ are commensurable values their ratios on the input and on the output for ordinary radar without SDTs are approximately identical.

Let us determine, by how many times it is necessary to increase the density of the dipole reflectors in a cloud or what is the same thing, increase the ratio $\frac{P_{\text{п}}}{P_{\text{с}}}$ on the input in order to obtain on the output of the alternate-period compensation circuit of the radar with SDTs the same ratio $k_{\text{в}}=\left(\frac{P_{\text{п}}}{P_{\text{с}}}\right)_{\text{вых}}$, as in the case of ordinary pulse radar. Naturally, it is assumed that the terminal devices in both cases are identical.

In deriving the formula all the basic physical peculiarities of the examined principle of discriminating moving targets are considered and at the same time a number of assumptions is made, simplifying the solution of the problem without the loss of physical clarity.

We will consider that on the input of the radar there acts either one effective signal, or only an interference signal. In other words, we will not consider the joint passage of interference and effective signals.

If one were to record the effective signal on the input of the C_2 mixer in the form of a harmonic vibration

$$u_{\text{с}}(t) = u_{\text{с}} \cos(\omega_0 t + \varphi), \tag{7.38}$$

where $u_{\text{с}} = \sqrt{2P_{\text{с вх}}}$, and $P_{\text{с вх}}$ – the power of the effective signal on the input, then the resultant oscillation on the input of the i-f amplifier due to the coherence of the accepted oscillation and the CO oscillator has the form

$$u_{\text{вх}} = \text{К}_{\text{пр}}\left[u_{\text{г}} \cos \omega_{\text{пр}} t + u_{\text{с}} \cos(\omega_{\text{пр}} t + \varphi)\right],$$

where $\text{К}_{\text{пр}}$ – a constant coefficient.

Assuming the linearity of the detector, we will record the envelope on the output of the video amplifier

$$u = \text{К}_1 \sqrt{u_{\text{г}}^2 + u_{\text{с}}^2 + 2u_{\text{с}} u_{\text{г}} \cos\varphi}, \tag{7.39}$$

where К_1 – a constant coefficient.

Since $u_{\text{г}} \gg u_{\text{с}}$, then

$$u \approx \text{К}_1 u_{\text{г}}\left(1 + \frac{u_{\text{с}}}{u_{\text{г}}} \cos\varphi\right). \tag{7.40}$$

Accordingly for the interference signal

$$u'_{\text{п}} \approx \text{К}_1 u_{\text{г}}\left[1 + \frac{u_{\text{п}}(t)}{u_{\text{г}}} \cos\varphi_{\text{п}}(t)\right]. \tag{7.41}$$

If the radial component of the target speed in the direction to a given radar is equal to v_r, then the lead of the phases ϕ for the pulse repetition period (T) will be

$$\varphi = \omega_{д} T = 2\pi \frac{2v_r}{c} f_0 T = 2\pi \frac{2v_r}{\lambda} T.$$

Here $\omega_{д} = 2\pi f_{д}$ – a Doppler frequency.

The lead of the phases ϕ after n repetition periods (after sending n pulses)

$$\varphi_n = (n - 1)\omega_{д} T, \tag{7.42}$$

where n = 1, 2, 3, ...

The difference in the voltages of u(n) and u(n + 1) of two adjacent pulses of n-th and (n + 1)-th effective signal is equal to

$$\Delta u_c(n) = u(n) - u(n + 1)$$

or, considering u_c = const, taking into account (7.42),

$$\Delta u_c(n) = к_{пр} u_c [\cos(n - 1)\omega_{д} T - \cos n\omega_{д} T]. \tag{7.43}$$

Here $к_{пр}$ – the transfer function of the whole processing circuit of the receiver-indicator channel (to the subtractor);

$$\Delta u_c(n) = к_{пр} 2u_c \sin\frac{\omega_{д} T}{2} \sin\frac{2n - 1}{2} \omega_{д} T. \tag{7.44}$$

On the output of the difference circuit the pulse envelope is a harmonic oscillation of frequency $\omega_{д}$ and amplitude

$$\Delta u_c = к_{пр} 2u_c \sin\frac{\omega_{д} T}{2}. \tag{7.45}$$

In the SDTs circuit with double subtraction the output voltage is determined as the difference

$$\Delta^{(2)} u_c(n) = \Delta u_c(n) - \Delta u_c(n + 1). \tag{7.46}$$

With the help of (7.44) we obtain

$$\Delta^{(2)} u_c(n) = к_{пр} 4u_c \sin^2\frac{\omega_{д} T}{2} \cos n\omega_{д} T. \tag{7.47}$$

The amplitude of oscillation on the output of the double subtraction circuit is equal to

$$\Delta^{(2)}u_{c} = \text{к}_{\text{пр}} 4u_{c} \sin^{2}\frac{\omega_{\text{д}}T}{2}. \qquad (7.48)$$

Accordingly with m-fold subtraction the amplitude of the oscillations on the output is

$$\Delta^{(m)}u_{c} = \text{к}_{\text{пр}} u_{c} 2^{m} \sin^{m}\frac{\omega_{\text{д}}T}{2}. \qquad (7.49)$$

The alternate-period difference of the interference signals taking into account (7.41) will be recorded in the following way:

$$\Delta^{(1)}u_{\text{п}} = \text{к}_{\text{пр}} [u_{\text{п}}(t)\cos\varphi_{\text{п}}(t) - u_{\text{п}}(t+T)\cos\varphi_{\text{п}}(t+T)]. \qquad (7.50)$$

The subsequent conversions (7.50) in principle can be carried out in two ways.

The first way (more valid) is based on the temporal representations of the interference signals. For its realization knowledge of the correlation function of the interference signal on the output of the envelope detector is necessary.

The alternate way based on the simplified frequency representation of the interference signals makes it possible to find a comparatively simple solution, but, naturally, very approximate.

Let us in the beginning use the first more valid way, having made, however, a number of assumptions with respect to the character of the movement of the dipoles in the cloud. We will consider that the process of turbulent diffusion develops comparatively slowly, so that the interference signal can be assumed steady for a time commensurable with the pulse repetition period of the suppressed radar.

Further let us approximate the power spectrum of the interference signal on the output of the envelope detector by the function

$$G(F)=\exp\left[-0{,}7\left(\frac{F}{F_{0{,}5}}\right)^{2}\right]. \tag{7.51}$$

Here $F_{0.5}$ – the spectrum width of the interference signal with respect to the half value of the standardized spectral density.

Let us record the difference of the interference signals (7.50) applying simpler designations,

$$\Delta^{(1)}u_{\text{п}}(t)=\text{к}_{\text{пр}}[u'_{\text{п}}(t)-u'_{\text{п}}(t+T)], \tag{7.52}$$

where

$$u'_{\text{п}}(t+T)=u_{\text{п}}(t+T)\cos\varphi_{\text{п}}(t+T);$$
$$u'_{\text{п}}(t)=u_{\text{п}}(t)\cos\varphi_{\text{п}}(t).$$

The power of the interference signal on the output of a single subtraction circuit is obtained as a result of averaging the square $\Delta^{(1)}u_{\text{п}}$:

$$P'_{\text{п вых}}=\overline{[\Delta^{(1)}u_{\text{п}}]^{2}}=\text{к}^{2}_{\text{пр}}\,\overline{[u'_{\text{п}}(t)-u'_{\text{п}}(t+T)]^{2}}$$

or

$$P'_{\text{п вых}}=\text{к}^{2}_{\text{пр}}\,2P_{\text{п вх}}[1-r(T)]. \tag{7.53}$$

Here $P_{\text{п вх}}$ is the power of the interference signal on the input of the receiver with single subtraction; r(T) is the standardized correlation function of interference signal on the output of the envelope detector calculated for the value $\tau = T$.

Due to the assumed steady state of the interference signal, the standardized correlation function can be presented in the following manner

$$r(\tau)=\frac{\overline{u_{\text{п}}(t)\,u_{\text{п}}(t+\tau)}}{\overline{[u_{\text{п}}(t)]^{2}}},$$

the vinculum above signifies the operation of averaging in time. With the help of (7.45) and (7.53) we determine the ratio of the powers of the interference and effective signals on the output of the single subtraction circuit

$$k_n^{(1)} = \left(\frac{P_n}{P_c}\right)^{(1)}_{\text{вых}} = \left(\frac{P_n}{P_c}\right)_{\text{вх}} \frac{1-r(T)}{2\sin^2\frac{\omega_c T}{2}}.$$

Let us assume that $k'_n = k_{n\,min}$ is the minimum necessary ratio of the powers of the interference and effective signals on the input of the indicqtor, at which the probability of correct target detection against a background of a cloud of passive interferences is less than a certain given value.

The value $k_{n\,min}$ corresponds to a certain value of the signal ratio on the input $k = \left(\frac{P_n}{P_c}\right)_{\text{вх}}$, which for given conditions will also be the minimum necessary from the point of view of providing suppression of the radar. In other words, the coefficient of suppression of radar with single subtraction is equal to

$$k_{\text{п}} = k_{n\,min} \frac{2\sin^2\frac{\omega_c T}{2}}{1-r(T)}. \qquad (7.54)$$

The standardized correlation function $r(\tau)$ is determined with help of Khinchine conversion, if the spectral density $G(\Omega)$ of the random process is known. Using the earlier indicated approximation (7.51), for the spectral density of the interference signal on the output of the envelope detector we obtain

$$r(\tau) = \eta \int_0^\infty G(\Omega)\cos\Omega\tau\, d\Omega.$$

Here η is the normalizing factor,

$$r(\tau) = \eta \int_0^\infty e^{-0.7\frac{\Omega^2}{\Delta\omega_{0.5}^2}} \cos\Omega\tau\, d\Omega \qquad (7.55`$$

or

$$r(\tau) = e^{-\frac{\omega_{0.5}^2\tau^2}{0.7}}. \qquad (7.56$$

Formula (7.54) determines the value of the coefficient of suppression assuming the ideal operation of the moving-target selection [SDTs] circuit. Actually the quality of operation of the SDTs circuit to a considerable extent will be affected by the instability of the transmitter of the local oscillator, the coherent oscillator, the frequency of pulse repetition, imperfectness of the subtraction operation itself, and the characteristic motion of the antenna. All of this taken together can be considered an equivalent expansion of the interference signal spectrum. Within a certain margin one may assume that the equivalent width of the spectrum on the average is increased by not less than twice in the centimeter range and by 1.4-1.5 times in the decimeter range.

A more exact calculation of the instability of the elements of a pulse-coherent radar circuit can be carried out not by increasing the width of the spectrum of signals reflected from the cloud by an appropriate number of times, but by the additive addition to the spectrum of the interference signal of a certain equivalent spectrum allowing for a different type of instability. However calculating the instabilities by a more exact method somewhat complicates the calculation formulas and does not increase by any significant extent the accuracy of the calculation on the whole.

The minimum necessary ratio of powers of interference and effective signals on the input of the indicator for indicators with brightness marking is $k_{\text{в мин}} = 1{,}5 \div 2$ [60].

Taking into account what has been said we will obtain the following approximate formulas for calculating the coefficient of suppression of radar with single subtraction:

a) the centimeter wave range

$$k_{\text{п}} = \frac{3\sin^2\left(\frac{2\pi v_r}{\lambda}T\right)}{1-e^{-8(F_{\text{н}}T)^2}}; \qquad (7.57)$$

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$$k_{\text{п}} = \frac{3\sin^2\left(\frac{2\pi v_r}{\lambda} T\right)}{1 - e^{-28(F_{\text{п}}T)^2}}. \tag{7.58}$$

Here $F_{\text{п}}$ is the equivalent width of the power spectrum of the interference signal.

With sufficient accuracy it is possible to consider $F_{\text{п}} = F_{0,5}$. The value of $F_{\text{п}}$ can be estimated by the formulas of 7.4 on the basis of meteorological data.

In Fig. 7.19 in accordance with (7.57) the coefficients of suppression are plotted in a logarithmic scale depending upon $\frac{v_r}{\lambda} T$. The product of $F_{\text{п}}T$ is the parameter of the plotted family of curves. In view of the periodicity of function

$$k_{\text{п}} = k_{\text{п}}\left(\frac{v_r}{\lambda} T\right)$$

the graphs were plotted for the values of the arguments $0 < \frac{v_r}{\lambda} T < 1.$ The transition to other values of the argument is easily accomplished by the elimination of its values, which are multiples of 2π.

As can be seen from the given dependences, the value of the coefficient of suppression essentially depends on the spectrum width of the interference signal, the pulse repetition period, the radial component of speed and wavelength.

The ratio of the powers of interference and effective signals on the output of an SDTs circuit with single subtraction in accordance with (7.43) will equal;

a) Centimeter range radar

$$k_{\text{в}} = \left(\frac{P_{\text{п}}}{P_{\text{c}}}\right)_{\text{вх}} \frac{1 - e^{-56(F_{\text{п}}T)^2}}{2\sin^2\left(\frac{2\pi v_r}{\lambda} T\right)}; \tag{7.59}$$

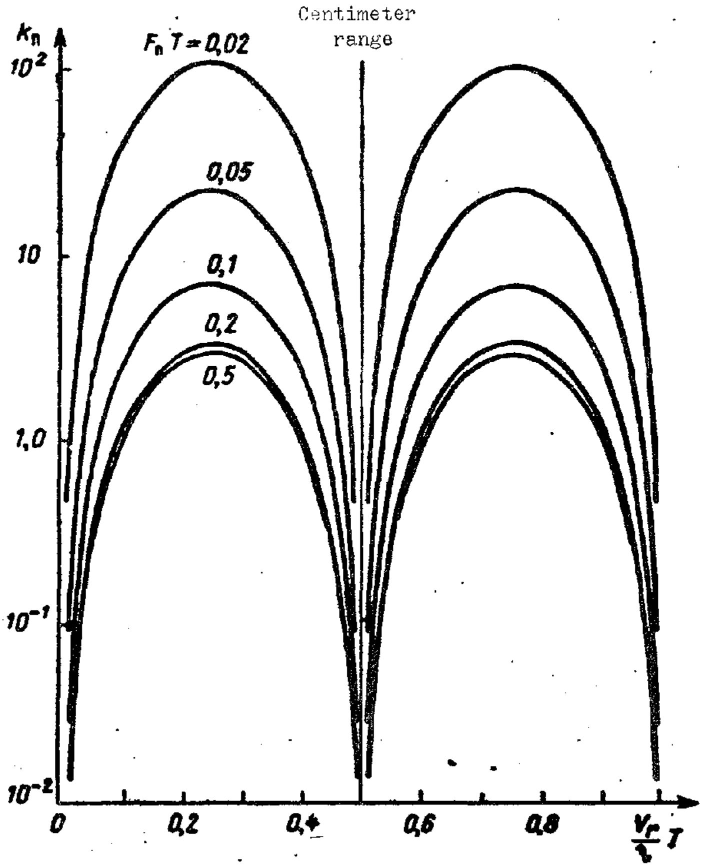

Fig. 7.19. The dependence of the coefficient of suppression by passive interferences of pulse-coherent radar in the centimeter range on the parameter $\frac{v_r}{\lambda}T$.

b) Meter range radar

$$k_{в} = \left(\frac{P_{п}}{P_{с}}\right)_{вх} \frac{1 - e^{-28(F_{в}T)^2}}{2\sin^2\left(\frac{2\pi v_r}{\lambda}T\right)}. \tag{7.60}$$

The dependences $k_{в} = k_{в}v_rT/\lambda$ are plotted in Fig. 7.20 assuming $k = (P_{п}/P_{c})_{вх} = 1$. In order to obtain with their help the values of $k_{в}$, corresponding to the other values of k, it is necessary to increase the ordinate of the curves by k times.

The application of double subtraction in radar somewhat increases the necessary expenditure rates of dipole reflectors. The power of

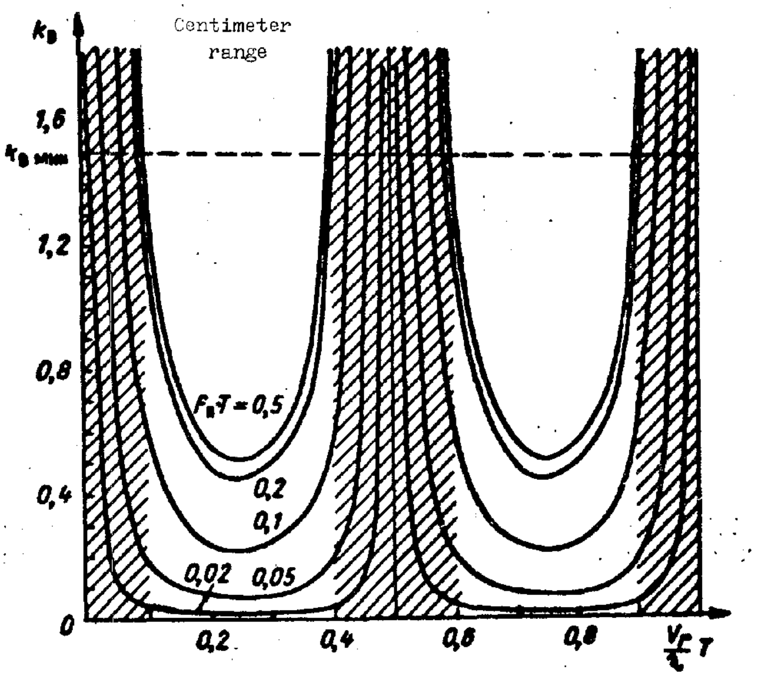

Fig. 7.20. The dependence of the interference ratio on the output of the SDTs circuit with single subtraction on the parameter $\frac{v_r}{\lambda}T$.

the interference signal on the output of the double subtraction circuit will be equal to

$$P_{п}^{(2)} = \overline{[\Delta^{(2)} u_{п}(t)]^{2}}.$$

Due to the principle of functioning of the double subtraction circuit

$$\Delta^{(2)} u_{п}(t) = \Delta^{(1)} u_{п}(t) - \Delta^{(1)} u_{п}(t+T), \qquad (7.61)$$

consequently,

$$(P_{п}^{(2)})_{вых} = 2P'_{п\,вых} - \overline{2\Delta^{(1)} u_{п}(t)\,\Delta^{(1)} u_{п}(t+T)}. \qquad (7.62)$$

Hence by analogy with (7.53) we find

$$(P_{п}^{(2)})_{вых} = 4P_{п\,вх}[1 - r(T)][1 - r^{(2)}(T)]. \qquad (7.63)$$

Here $r^{(2)}(T)$ is the standardized correlation function of the random process $\Delta^{(1)}\dot{u}(t)$, calculated for the value $\tau = T$.

Accordingly with m-th subtraction the output power $(P^{(m)}_{\text{п}})_{\text{вых}}$ is equal to

$$P^{(m)}_{\text{п вых}} = \overline{[\Delta^{(m)}u(t)]^2} = 2^m P_{\text{п вх}} \prod_{j=1}^{m} [1 - r^{(j)}(T)], \tag{7.64}$$

where $\prod\limits_{j=1}^{m}$ is the sign of the product.

The coefficient of correlation in circuits with double subtraction $r^{(2)}(T)$ can be calculated by comparing (7.62) and (7.63). Direct calculation of the correlation function of the random process $\Delta^{(1)}u_{\text{п}}(t)$ gives

$$\begin{gathered}\overline{\Delta^{(1)}u_{\text{п}}(t)\,\Delta^{(1)}u_{\text{п}}(t+\tau)} = \\ = P_{\text{п вх}}[2r(\tau) - r(\tau - T) + r(\tau + T)],\end{gathered} \tag{7.65}$$

whence it follows that

$$r^{(2)}(T) = \frac{2r(T) - r(2T) - 1}{2[1 - r(T)]}. \tag{7.66}$$

Substituting (7.66) in (7.63), we find

$$P^{(2)}_{\text{вых}} = 2P_{\text{п вх}}[3 - 4r(T) + r(2T)]. \tag{7.67}$$

Accordingly the ratio of the powers of the interference and effective signals on the output of the SDTs circuit with double subtraction is

$$\left(\frac{P_{\text{п}}}{P_{\text{с}}}\right)^{(2)} = k_{\text{п}}^{(2)} = \left(\frac{P_{\text{п}}}{P_{\text{с}}}\right)_{\text{вых}} \frac{3 - 4r(T) + r(2T)}{8\sin^4 \frac{\omega_{\text{д}} T}{2}}, \tag{7.68}$$

where r(T) and r(2T) are determined by formula (7.56).

Taking into account the instabilities of the SDTs circuit itself respectively for radar of meter and centimeters of ranges it is possible to record

$$r(T)=e^{-56(TF_{\text{п}})^2},\qquad (7.69)$$
$$r(T)=e^{-224(TF_{\text{п}})^2}.$$

The coefficient of suppression of radar with double subtraction is equal to

$$k_{\text{п}}=k_{\text{в мин}}\frac{8\sin^4\frac{\omega_{\text{д}}T}{2}}{3-4r(T)+r(2T)}.\qquad (7.70)$$

Usually $k_{\text{в мин}}$ is equal to 1.5.

It is necessary to note that the formulas obtained are approximate, however for practical purposes their accuracy can be sufficient. Actually, the initial data relative to the effective area of scattering of a pack can vary for one and the same type of dipole by 1.5-2 times (due to the variation of parameters of dispersion, increased adhesion of the dipoles as a result of prolonged storage in unsuitable locations, increased breakage due to defective packing and arrangement in containers and so forth). All of this justifies the use of rough estimates. Moreover, approximate, but sufficiently appropriate physical investigations make it possible to develop new possibilities for increasing the effectiveness of dipole reflectors as jamming agents.

In conclusion let us dwell on the very approximate formula for the coefficient of suppression based on the simplified frequency representation of interference signals.

Let us consider that the power spectrum of the interference signal is known to us (Fig. 7.21). Let us plot a rectangular spectrum equivalent to it in power. The maximum frequency of the equivalent rectangular spectrum $\Omega_{\text{п}}$ will be determined from the condition

$$\Omega_{\text{п}}=\frac{1}{G_{\text{макс}}}\int_0^\infty G(\Omega)\,d\Omega.\qquad (7.71)$$

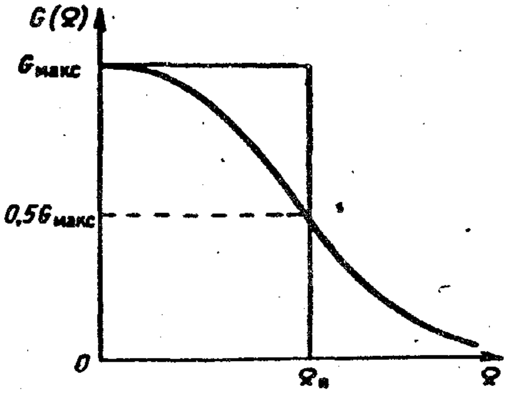

Fig. 7.21. The power spectrum of an interference signal created by passive interferences.

In the first approximation instead of $\Omega_п$ it is possible to take $\Omega_{0.5}$ – the spectrum width at a level of half spectral density.

Due to the rectangular spectrum

$$u_п(t) = u_п = \sqrt{2G_{макс}\Omega_п} = \sqrt{2P_{п\,вх}}.$$

Here $P_{п\,вх}$ is the power of the interference signal on the input of the receiver.

Let us replace the obtained rectangular continuous spectrum with harmonics of a discrete spectrum with frequencies respectively equal to $\frac{\Omega_п}{N}$, $2\frac{\Omega_п}{N}$, $\ldots, \Omega_п$. A rather large number N can be selected. Then

$$\cos\varphi_п(t) = \sum_{k=1}^{N} \cos k\,\frac{\Omega_п}{N}\,t. \tag{7.72}$$

Using the complex representation $\cos k\cdot\frac{\Omega_п}{N}\,t$ and applying the formula of a geometric progression, we find

$$\cos\varphi_п(t) = \cos\left(\frac{N+1}{2}\,\frac{1}{N}\,\Omega_п t\right)\frac{\sin\left(\frac{N}{2}\,\frac{1}{N}\,\Omega_п t\right)}{\sin\frac{\Omega_п t}{2}}. \tag{7.73}$$

If N is sufficiently large, then

$$\cos\varphi_{\text{п}}(t)\approx\cos\frac{\Omega_{\text{п}}}{2}t, \tag{7.74}$$

i.e., instead of N harmonics it is possible to consider one harmonic, the frequency of which coincides with the average frequency of the spectrum.

Consequently, expression (7.50) neglecting the instabilities of the SDTs circuit it is possible to record in this manner:

$$\Delta u_{\text{п}}=\kappa_{\text{пр}}u_{\text{п}}\left[\cos\frac{\Omega_{\text{п}}}{2}t-\cos\frac{\Omega_{\text{п}}}{2}(t+\tau)\right] \tag{7.75}$$

or

$$\Delta u_{\text{п}}=2\kappa_{\text{пр}}u_{\text{п}}\sin\frac{\Omega_{\text{п}}T}{4}\sin\frac{\Omega_{\text{п}}}{2}\left(t+\frac{T}{2}\right). \tag{7.76}$$

The amplitude of the interference oscillation was equal to $2\kappa_{\text{пр}}u_{\text{п}}\sin\Omega_{\text{п}}T/4$, hence the ratio (which interests us) of the power of the interference signal to the power of the effective signal on the output of the single subtraction circuit was equal to

$$k_{\text{в}}=\left(\frac{P_{\text{п}}}{P_{\text{c}}}\right)_{\text{вх}}\frac{\sin^2\frac{\Omega_{\text{п}}T}{4}}{\sin^2\frac{\omega_{\text{д}}T}{2}}. \tag{7.77}$$

Usually $\Omega_{\text{п}}T\ll 1$, therefore

$$k_{\text{в}}=\left(\frac{P_{\text{п}}}{P_{\text{c}}}\right)_{\text{вх}}\frac{\Omega_{\text{п}}^2T^2}{4^2\sin^2\frac{\omega_{\text{д}}T}{2}}.$$

In SDTs circuits with m-th subtraction the amplitude of the interference signal on the output $\Delta^{(m)}u_{\text{п}}$ will be equal to

$$\Delta^{(m)}u_{\text{п}}=\kappa_m2^m\sin^m\frac{\Omega_{\text{п}}}{4}T\approx\kappa_m2^m\frac{(\Omega_{\text{п}}T)^m}{4^m}, \tag{7.78}$$

where κ_m is the proportionality factor.

Accordingly the ratio of the powers of the interference and effective signals was equal to

$$k_{\text{в}}^{(m)} = \left(\frac{P_{\text{п}}}{P_c}\right)_{\text{вх}} \frac{(\Omega_{\text{п}} T)^{2m}}{4^{2m} \sin^{2m} \frac{\omega_{\text{д}} T}{2}}. \qquad (7.79)$$

Hence a formula is obtained for the coefficient of suppression with m-th subtraction

$$k_{\text{п}}^{(m)} = k_{\text{в мин}} \frac{4}{(2\pi)^{2m}} \frac{\sin^{2m}\left(\frac{2\pi v_r}{\lambda} T\right)}{(F_{\text{п}} T)^{2m}}. \qquad (7.80)$$

Accordingly for single subtraction

$$k_{\text{п}}^{(1)} = k_{\text{в мин}} \frac{0{,}4 \sin^2 \frac{2\pi v_r}{\lambda} T}{F_{\text{п}}^2 T^2}. \qquad (7.81)$$

Reliable suppression of the effective signal on indicators with brightness markers arises when $k_{\text{в мин}} = 1{,}5 \div 2$, respectively:

$$k_{\text{п}}^{(1)} = (0.6 \div 0.8) \frac{\sin^2 \frac{2\pi v_r T}{\lambda}}{F_{\text{п}}^2 T^2}. \qquad (7.82)$$

Calculation of the instabilities can be approximately carried out with an appropriate increase of $\Omega_{\text{п}}$ – the spectrum width of the interference signal analogous to what was done earlier.

Formulas (7.80), (7.81), (7.82) due to the rough approximation of the continuous spectrum of the interference signal give several times higher values for the coefficient of suppression than corresponding formulas (7.54), (7.57), (7.58) and (7.70).

CHAPTER 8

FALSE TARGETS AND RADAR TRAPS

8.1. Introduction

One method of jamming is the creation of false information in the target-distribution, guidance and homing guidance contours. The simplest way from a theoretical viewpoint that this problem can be solved is by the application of radar traps or false targets launched from aircrafts or from earth to overcome the enemy antiaircraft.

False targets create on radar screen contours of target-distribution markers similar to the markers from real targets. This to a considerable degree hampers the identification of real targets by the operator (or machine). In individual cases due to the limitation of observation time the distinguishing of true targets against the background of false targets is impossible which forces the enemy to act both against the false targets, as well as against the other targets.

The large number of false targets along with the disorientation of the radar operators overloads the data processing system. As a result it reduces the number of possible attacks or launches of AA guided missiles against the protected aircraft.

As false targets there are used rockets equipped with booster or sustainer engines, the presence of which makes it possible to carry out autonomous, controlled or free flight for a prolonged time period (up to several tens of minutes). So that the false-target rocket creates a signal with the same intensity and spectrum as the protected aircraft, it is equipped with appropriate means – active and passive retranslators.

An example of a false target is the GAM-72 rocket (the United States) which was included in the armament of B-52 and B-47 strategic bombers in 1960 [63]. On the rocket is mounted an engine with a thrust of 1100 kg. The initial weight of the rocket is 500 kg; the body length is 4 m; the wingspan is 1.6 m. The flying range is 360 km and the ceiling is about 15,000 m.

A GAM-72 false target can create on radar screens markers analogous to markers from true targets. Furthermore, on rocket there is mounted equipment for jamming radiotechnical, acoustical and infrared means of detection and guidance. A B-52 aircraft can carry several such rockets. Depending upon the assignment of the false target its guidance can be carried out by radio, or autonomously by a preset program.

The false target identification vector should contain at least three basic components: α_1 – the amplitude of the signal reflected from the false target; α_2 – speed; α_3 – acceleration.

When necessary the dimension of the false target identification vector can be increased.

In contrast to false targets radar traps are designed to disrupt automatic target tracking by radar or by a homing device. They provide the possibility of switching the automatic tracking contour from the true target to the trap.

8.2. The Application of False Targets in Target-Distribution Contours

The basic problems in using false targets in target-distribution contours are:

– the disorientation of radar operators and overloading the contour computers (system of data processing);

– increasing the time necessary to identify the shape of the target (determining the true targets);

– to divert the strike forces of antiaircraft defense (fighters, rockets) to destroy the false targets.

The effectiveness of false target application depends on the ratio of the total number of targets (real and false) to the strike forces of the antiaircraft defense (fighters, guided antiaircraft rockets, ZUR), and also on the tactics applied by the conflicting sides. The effectiveness of false targets can be estimated by decreasing the probability of the destruction of the covered aircraft.

In the case of mass application of false targets the probability of the destruction of an aircraft covered by a group of false targets can be calculated by the formula

$$P_m(n)=\frac{m}{n}P, \tag{8.1}$$

where n is the total amount of targets (false and real) in the group; m is the number rockets launched; P is the probability of destruction of a target in one shot.

Formula (8.1) is valid with the following assumptions:

– $m \le n$;

– the selection of the targets (false or real) by the target-distribution system to be fired on is equiprobable;

– for each target there is one rocket launching (one fighter attack) regardless of whether the target is false or real.

The dependence of the probability of the destruction of one true target $P_m(n)$ covered by n - 1 false targets after m shots is shown in Fig. 8.1. The curves are plotted for the probability of target destruction by one rocket P = 0.8.

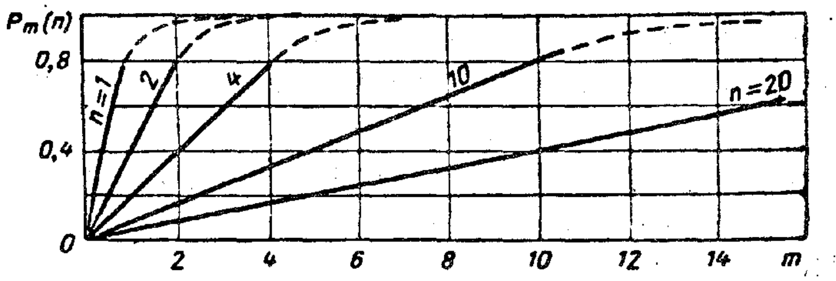

Fig. 8.1. The dependence of the probability of the destruction of one true target $P_m(n)$ covered by n - 1 false targets after m shots (launchings).

It is easy to see that for the indicated conditions the use of false targets essentially lowers the probability of aircraft destruction. Thus, in the case of the coverage of an aircraft by one false target (n = 2) the probability of its destruction by one rocket (m = 1) is halved [$P_m(n) = 0.4$] as compared to the probability of the destruction of an uncovered aircraft (n = 1). However the reduction of the effectiveness of antiaircraft defense in practice to zero [$P_m(n) \approx 0.05-0.1$] is provided by the application of a relatively large number of false targets (10-20). This is one of the deficiencies of the investigated scheme of using false targets in target-distribution contours.

The reproduction of the amplitude component of the false target identification vector is ensured either by amplifiers-retranslators, or by passive reradiators of electromagnetic energy.

Active amplifiers-retranslators can be effective in the meter and decimeter wave ranges and for large distances to suppressed radar in connection with their limited power possibilities. At short distances the signal power created by a retranslator on the input of the suppressed device will be less than the power of the effective signal reflected from the covered aircraft, in consequence of which the operator can identify a real target in the midst of false targets.

Passive reradiators (various kinds reflectors) make it possible to obtain a sufficiently large effective trap scatter area [ESA] commensurable with the ESA of a covered aircraft in the centimeter wave range.

In individual cases on rockets-traps jammers and devices for dropping of dipole reflectors can be mounted.

8.3. The Application of Radar Traps in Guidance and Homing Guidance Contours

The application of traps in a guidance or homing guidance contour should lead, as a rule, to its locking-on the false target. The time of locking-on should be commensurable with average time of guidance (homing guidance) of the destructive weapons of antiaircraft defense. The launching (release) of a trap in this case should be carried out after the locking-on of the guidance (homing guidance) contour on the real target. The successful application of a trap leads to the disruption of the AA guided missile (fighter) attack or to the missing of the covered aircraft by same.

The identification vector of the trap should have the components ensuring its lock-on, taking into account the amplitude (power) characteristics, speed and acceleration. The interference signal generated by the trap on the input of suppressed automatic tracking system should exceed the effective signal, in order to ensure the possibility of the switching of the simplest servosystems onto the trap.

Let us note that the discussion of radar traps conducted below assumes the presence in the suppressed agent of a simple system of automatic guidance, changing the variable parameters when the input value is deflected from a certain given value, usually zero. In the case of guidance systems of the more complicated type, having self-adjusting elements for optimum conditions in accordance with the information, coming from search and analyzing devices, the most perfected trap under the best conditions can ensure the locking-on of the guidance or homing guidance contour on itself with a probability approximately equal to 0.5.

According to the method of combat application radar traps can be categorized into guided, towed and dropped traps.

Guided Traps

Guided traps, like false targets applied in target-distribution contours are rockets with passive and active reradiators of electromagnetic energy. On the rocket-traps there can be mounted both booster and also sustainer engines providing guided flight (by radio or by program) for a period of time of from several seconds to several minutes.

Rocket-traps ensure the disruption of guidance (homing guidance) by decoying to itself the attacking rocket (or fighter).

Direction of launch of the rocket-trap is determined by direction of attack and by the relationship of the velocity vectors of the target, trap and attacking missile. For the successful application of a rocket-trap a covered aircraft should simultaneously with the launching of the trap change its speed and direction.

The initial speed of the rocket-trap is determined by the dynamic characteristics of the servosystems (for angle, speed, distance) of the guidance (homing guidance) contour. In the first instant after the dropping of the trap its speed of departure from the carrier aircraft should ensure the decoying of all the gates of the servomechanisms into the false target, otherwise the application of the trap would be futile. Tentatively the initial speed of the trap should be selected from the condition nonresolution of the covered aircraft and trap by angle, distance and speed in the first moment of time. The covered aircraft and trap can be resolved from the accelerations (overloads) which it is necessary to consider in setting up jamming.

Let us find the equation of jamming for the case of application of rocket-traps in the homing guidance contour of controlled antiaircraft rockets ZUR using method of semiactive homing (Fig. 8.2).

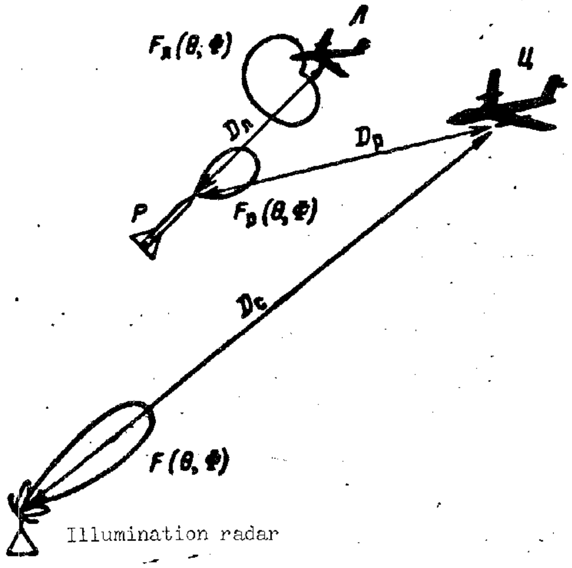

Fig. 8.2. A variant of the application of rocket-traps in a homing guidance contour of controlled antiaircraft rockets ZUR.

Let us assume that the ground radar "illuminates" target (Ц), and the signals reflected from target Ц, are received by the head of the homing rocket P.

Let us assume that to disrupt the homing guidance of the rocket (P) the target aircraft launches a rocket-trap with an active reradiator of the illumination radar signals.

The power of the effective signal (the illumination radar signal) on the input of the receiver of the rocket homing device is equal to

$$P_{\text{с вх}} = \frac{P_c G_c F^2(\theta_c, \Phi_c)\,\sigma_{\text{ц}}}{4\pi D_c^2 4\pi D_p^2} A_r F_p^2(\theta'_c, \Phi'_c), \tag{8.2}$$

where $P_c G_c$ is the power potential of the illumination radar; $F(\theta, \Phi)$ is the standardized radiation pattern of the transmitting antenna of the illumination radar on the field; $F_p(\theta, \Phi)$ is the standardized radiation pattern of the receiving antenna of the rocket (P) homing device for the field; A_r is the maximum value of the effective area of the homing device receiving antenna; θ_c, Φ_c are the angular coordinates of the target aircraft (calculated from the maximum antenna radiation pattern $F(\theta, \Phi)$); θ'_c, Φ'_c are the angular coordinates of the target aircraft (calculated from the maximum antenna radiation pattern $F_p(\theta, \Phi)$); D_c is the distance between the target aircraft and the illumination radar; D_p is the distance between the attacking rocket and the target aircraft.

The power of the interference on the input of the rocket (P) homing device receiver is determined by relationship

$$P_{\text{п вх}} = \frac{P_{\text{п}} G_{\text{п}}}{4\pi D_{\text{л}}^2} A_r F_{\text{л}}^2(\theta_{\text{р}}, \Phi_{\text{р}}) F_{\text{р}}^2(\theta_{\text{л}}, \Phi_{\text{л}}) \gamma_{\text{п}} \frac{\Delta f_{\text{пр}}}{\Delta F_{\text{п}}}, \tag{8.3}$$

where $P_{\text{п}} G_{\text{п}}$ is the power potential of the retranslator (interference station) mounted onboard the rocket-trap; $F_{\text{л}}(\theta, \Phi)$ is the standardized radiation pattern of the transmitting antenna of the retranslator for the field; $\theta_{\text{р}}$, $\Phi_{\text{р}}$ are the angle coordinates of the rocket (P) calculated from the maximum antenna radiation pattern $F_{\text{л}}(\theta, \Phi)$; $\theta_{\text{л}}$, $\Phi_{\text{л}}$ are the angular coordinates of the trap calculated from the maximum antenna radiation pattern $F_{\text{р}}(\theta, \Phi)$; $D_{\text{л}}$ is the distance between the trap and attacking rocket; $\gamma_{\text{п}}$ is the coefficient considering the distinction of the polarizations of the antennas of the retranslator and the rocket homing device; $\Delta f_{\text{пр}}$ is the homing device receiver passband; $\Delta F_{\text{п}}$ is the width of the interference signal spectrum.

Using (8.2) and (8.3) we obtain the desired jamming equation

$$k = \left(\frac{P_{\text{п}}}{P_{\text{с}}}\right)_{\text{вх}} = \frac{P_{\text{п}} G_{\text{п}} 4\pi}{P_{\text{с}} G_{\text{с}} \sigma_{\text{ц}}} \frac{D_{\text{с}}^2 D_{\text{р}}^2}{D_{\text{л}}^2} \times$$

$$\times \frac{F_{\text{л}}^2(\theta_{\text{р}}, \Phi_{\text{р}}) F_{\text{р}}^2(\theta_{\text{л}}, \Phi_{\text{л}})}{F^2(\theta_{\text{с}}, \Phi_{\text{с}}) F_{\text{р}}^2(\theta'_{\text{с}}, \Phi'_{\text{с}})} \gamma_{\text{п}} \frac{\Delta f_{\text{пр}}}{\Delta F_{\text{п}}}. \tag{8.4}$$

Let us note that with the relaying of the signal it is possible to accept

$$\frac{\Delta f_{\text{пр}}}{\Delta F_{\text{п}}} = 1.$$

The value of output power $(P_{\text{п}})$ of an active or passive reradiator of a trap necessary to affect the basic lobe of the radiation pattern of the homing device antenna of the rocket can be found from the obtained jamming equation (8.4) by substituting in it the value of the coefficient of suppression $k_{\text{п}}$ instead of k. Let us find the power of a reradiator $P_{\text{п}}$, necessary to decoy an attacking rocket onto a trap at the time of launch. We will assume

$$F_{\text{л}}(\theta_{\text{р}}, \Phi_{\text{р}}) = F_{\text{р}}(\theta_{\text{л}}, \Phi_{\text{л}}) = F(\theta_{\text{c}}, \Phi_{\text{c}}) = F_{\text{р}}(\theta'_{\text{c}}, \Phi'_{\text{c}}) = 1,$$
$$D_{\text{л}} = D_{\text{р}}.$$

Taking into account the assumed condition from (8.4) we obtain

$$P_{\text{п}} = \frac{P_{\text{c}} G_{\text{c}} \sigma_{\text{ц}}}{4\pi D_{\text{c}}^2 G_{\text{п}}} k_{\text{п}}. \quad (8.5)$$

In connection with the fact that the purpose of the rocket-trap is to decoy onto itself the homing device antenna (the locking of the homing guidance contour onto the false target), the reflected signal simulated by the false target must exceed in power by several times the real reflected signal.

In setting up jamming to decoy fighters the use of traps presents more serious demands on retranslator power engineering than in the case of homing controlled antiaircraft rockets examined above. This first of all is connected with the necessity of providing the proper interference signal ratio for the comparatively small distances between the trap and the attacking fighter. The necessary power engineering for equipment onboard a trap to jam powerful pulse radar is sharply increased.

Towed Traps

Towed traps can be used to disrupt the attacks of rockets or fighters during the last stage of homing (homing guidance). These traps are towed by a bomber aircraft on a thin cable, the length of which can be several kilometers. In its packed state trap is located in a special compartment. Upon entering the most dangeous zones of antiaircraft defense it is launched with the help of a booster attachment.

The first experimentally applications of towed traps were carried out in the Second World War, when to reduce the effectiveness of German antiaircraft defense Anglo-American aviation used towed metallic nets as false targets. These nets creating powerful reflected signals attracted the gun-laying radar onto themselves.

The towed trap is equipped with amplifiers-retranslators and passive reradiators increasing the value of its effective scatter area up to the value of the ESA of the carrier aircraft. When necessary a jamming apparatus can be mounted on the trap.

Towed traps can be used for jamming guidance and homing guidance contours. Their effectiveness will be high, if in the initial moment of rocket (or fighter) guidance the covered aircraft and trap are represented as one target, i.e., they are not resolved by angles, distance and Doppler frequencies. The application of towed traps has a number of its own characteristics.

The distance of the towed trap from the aircraft is mainly determined by the resolving power of the suppressed system with respect to speed and angle.

The condition of the nonresolution of the aircraft and trap with respect to angle can be written in this form

$$L \leqslant \frac{\Delta\theta_p D}{\sin q}, \qquad (8.6)$$

where $\Delta\theta_p$ is the resolving power of the suppressed radar with respect to angle; D is the distance to the suppressed radar; q is the aspect of the trap.

The condition of the nonresolution of the aircraft and trap with respect to Doppler frequencies is determined by the bandwidth of the transmission of the "velocity gate" ΔF and by the difference of the Doppler frequencies of the trap and aircraft Δf:

$$\Delta f \leqslant \Delta F.$$

Here

$$\Delta f = \frac{2}{\lambda}\,(v_{\text{сл}} - v_{\text{л}}) = \frac{2\Delta v}{\lambda}, \qquad (8.7)$$

$v_{\text{сл}}$, $v_{\text{л}}$ are respectively the radial rates of closure of the attacking rocket with the covered aircraft and the trap (Fig. 8.3).

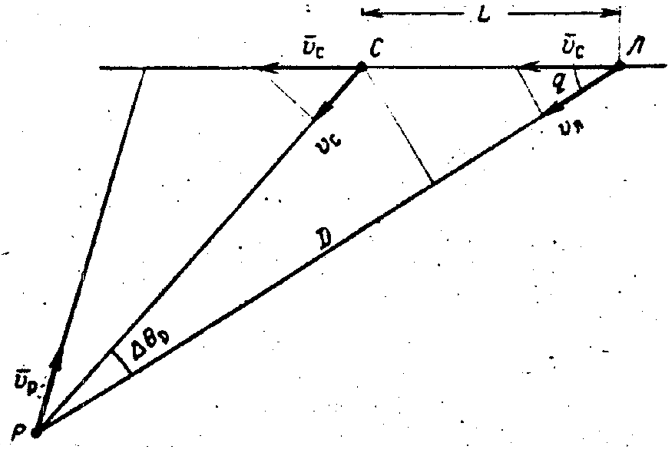

Fig. 8.3. A diagram of rocket (P) guidance to a target (C) towing a trap (Л).

Conditions (8.6) and (8.7) define the maximum permissible distance of the trap from the covered aircraft.

In Fig. 8.4 there are shown the zones of resolution of the towed trap and the aircraft with respect to angle I and the differences in the radial speeds (Doppler frequencies) II. In this same figure there is shown the shadow region of trap III by the aircraft. The aircraft shades the trap, if the attacking rocket is inside cone, the vertex angle of which is equal to

$$\alpha = \arcsin\frac{\kappa' R}{L},$$

where R is the effective destructive radius of the rocket; **к′** is the safety factor.

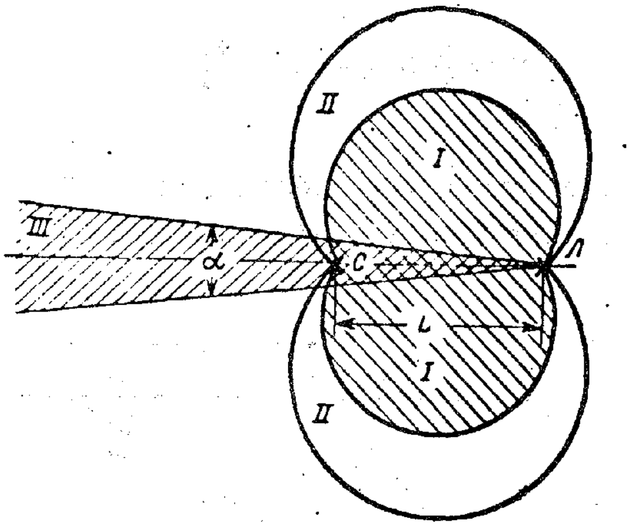

Fig. 8.4. The zones of resolution of the towed trap **(Л)** and the aircraft (C) with respect to angle (I), the difference in the radial speeds (II) and the shadow region (III) of the trap by the aircraft.

The value of zone III has considerable significance in attacks from the front hemisphere.

Dropped Traps

Dropped traps protect the aircraft from an attacking rocket (fighter). These traps do not have engines and are active or passive reradiators possessing a greater effective scatter area than the covered aircraft being subjected to attack. In the simplest case the trap can be a corner reflector or a pack of dipole reflectors.

A dropped trap can be locked-on by the servomechanism of the rocket (fighter), if the following conditions are fulfilled:

– the effective scatter area of the trap is larger than the ESA of the covered aircraft;

– the action time of the false target signals is greater than or equal to the time constant of the servomechanisms for angle, speed and distance.

The second condition puts rather stringent demands on the construction of a dropped trap.

The duration of the signal effect from the trap on the homing guidance system for pulse systems is determined by the time of the stay of the trap in the pulse volume, and for continuous radar – by the time of stay of the radial component of the relative velocity of the trap within the limits of the velocity passband of the servomechanism, and also by the time of stay of the trap within the limits of antenna radiation pattern of the suppressed radar.

Let us define the condition of the lock-on of the dropped trap by the automatic target tracking pulse radar.

In order to determine the time of stay of the trap and the target in the same pulse volume, it is necessary to calculate the trajectory of the false target dropped from an aircraft, when it is in its free fall. As is known from ballistics, the trajectory of a free fall is determined by the characteristic time θ, height and flying speed of the aircraft, from which the trap is dropped.[1]

In Fig. 8.5 there are given the approximate trajectories of a trap fall in a moving coordinate system connected with a bomber for different values of characteristic fall time $\theta_1 > \theta_2 > \theta_3$.

As can be seen from the figure, the characteristic fall time has an essential influence on the time of stay of the aircraft and the trap in the pulse volume. When $\theta = \theta_2$ the time of stay of a trap in the volume of a sphere with R = const is greater than when $\theta = \theta_1$.

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[1] The characteristic fall time is the fall time of a body from a height of 2000 m under the conditions of a standard atmosphere.

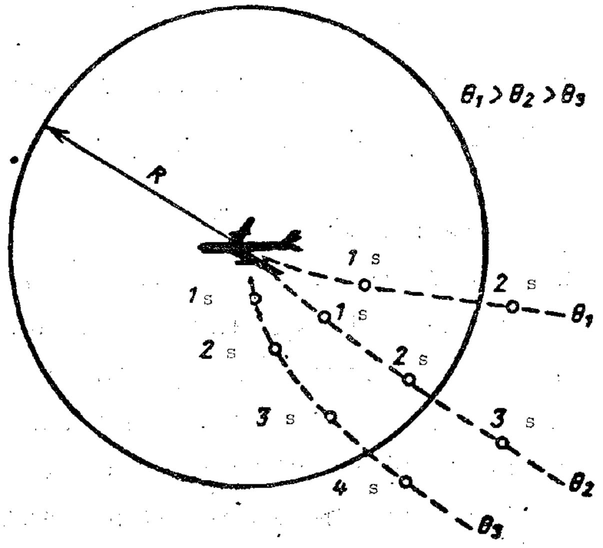


Fig. 8.5. The fall trajectory of a dropped trap without an engine.

The characteristic fall time of a trap is determined by the well-known formula [101] from ballistics

$$\theta = 20{,}2 + \frac{id^2}{G} C_x \, 10^3, \tag{8.8}$$

where $\frac{id^2}{G} C_x \, 10^3$ is the ballistic coefficient; i is the form factor of the trap; d is the diameter of the trap; G is the weight of the trap; C_x is the aerodynamic coefficient.

From formula (8.8) and Fig. 8.5 it follows that in order to increase the stay time of the trap in the pulse volume it is necessary to try to decrease its characteristic fall time θ.

In connection with the fact that the dimensions of the trap are determined by the dimensions of the reradiators mounted on it, a decrease in the characteristic time θ is attained by increasing the weight of the traps [see formula (8.8)]. Let us determine, for example, the weight of a trap, which is a Luneberg lens with a diameter d = 0.5 m ($\sigma \approx 190$ m^2, $\lambda = 5$ cm). Let us assume i = 8, $C_x = 0.2$. Then the necessary weight of the dropped traps will be approximately:

G = 500 kg, if θ = 21 s;
G = 70 kg, if θ = 26 s.

The given example shows that dropped traps must have considerable weight.

In order to increase the stay time of the dropped trap (with small weight) in a rather close proximity to the target, it is necessary to give it an initial speed in the direction of travel of the aircraft, i.e., to equip the trap with a booster. Accordingly its trajectories in a moving coordinate system connected with the bomber vary changed. In Fig. 8.6 there is given the trajectory of a trap equipped with a booster. Also in the figure the dotted line represents the trajectory of a trap not having a booster engine.

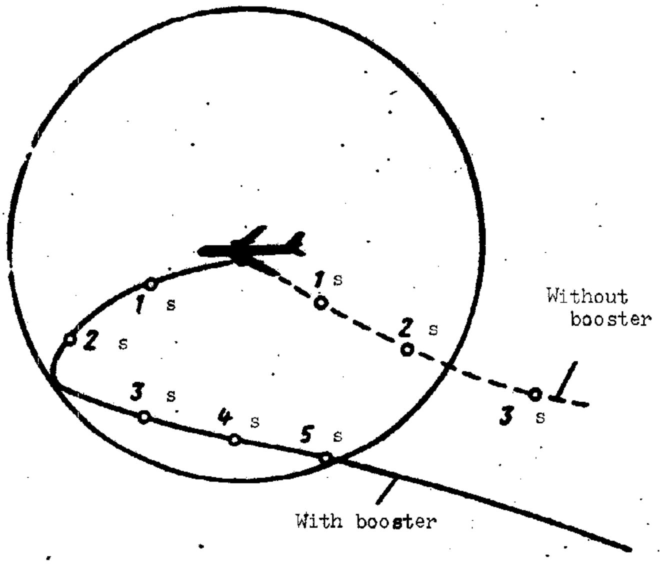


Fig. 8.6. The trajectory of travel of a trap equipped with a booster.

Until now we have assumed that there were no differences in the methods of creating interferences on guidance and homing guidance contours with the help of traps. However the application of traps for guidance contours, which are a servomechanism of the pulse type, has significant peculiarities. These peculiarities concern, mainly, maximum speed of separation of the trap from the covered aircraft, which in the latter case is determined by the discreteness of information about the target coordinates, which is received.

In the simplest guidance contour of the pulse type data about target position proceed with a certain off-duty factor, the value of which depends on the speed of rotation of the radar antenna (usually circular scan radar). The period of information arrived in contemporary systems is estimated by a value of the order of several seconds.

In launching a trap at the moment of irradiation of the covered aircraft, it is necessary to ensure the conditions (initial speed, average speed for the period of radar scan), under which the launched trap during one scan cycle could not emerge beyond the limits of the region near the covered aircraft, each point of which is accessible to this aircraft with respect to the conditions of overloads and possibilities of the power plant for the shown time. The absolute value of the initial speed in this case does not have a specific value. It is important, that the average speed for the scan period of the suppressed radar be lower than a certain value.

8.4. Methods of Increasing the Effective Scatter Area of False Targets

There are two basic methods of increasing the ESA of false targets:

- the application of retranslator amplifiers of received signals;
- the application of passive reradiators.

Retranslator Amplifiers

The technical realization of retranslator amplifiers does not cause fundamental difficulties. In Fig. 8.7 there is presented a block diagram of the simplest retranslator. The signals of the suppressed radar are received by the receiving antenna A_1, are amplified in the preamplifier PU and enter the input of the final power amplifier OU.

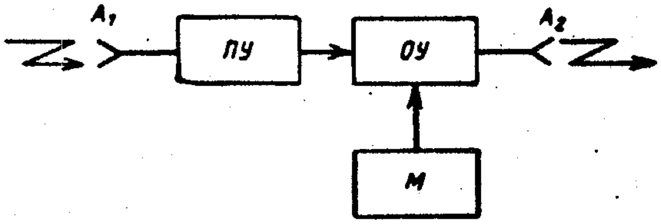


Fig. 8.7. Block-diagram of relay [retranslator] amplifier.

In the power amplifier the signals are modulated in amplitude by the noise voltage formed by modulator M. This is necessary for simulation of the fluctuations of the effective area of the real target. After appropriate amplification in the final amplifier OU the signal is radiated through antenna A_2.

Usually travelling-wave tubes are used as amplifiers; they have a wide passband and a high amplification factor.

Let us determine the amplification factor of the relay ensuring the necessary magnitude of interference signal power on the input of a given radar (Fig. 8.8).

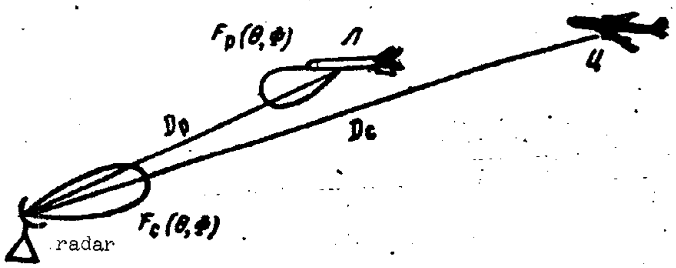


Fig. 8.8. A diagram of the application of a trap with a responder.

The power of the signal of the suppressed radar on the output of the relay receiving antenna is equal to

$$P_{\text{вх р}}=\frac{P_c G_c}{4\pi D_p^2}F_c^2(\theta_p, \Phi_p)F_p^2(\theta_c, \Phi_c)A_{r1}, \tag{8.9}$$

where $P_c G_c$ is the power potential of the suppressed radar; $F_c(\theta, \Phi)$ is a function describing the standardized antenna radiation pattern of the suppressed radar with respect to the field; $F_p(\theta, \Phi)$ is a function describing the radiation pattern of a relay receiving antenna with respect to the field; θ_p, Φ_p are the angular coordinates of the relay (trap) calculated from the maximum antenna radiation pattern of the suppressed radar; θ_c, Φ_c are the angular coordinates of the suppressed radar, calculated from the maximum antenna radiation pattern of the receiving antenna of the relay; A_{r1} is the maximum equivalent area of absorption of the relay receiving antenna; D_p is the distance between the relay (trap) and the suppressed radar.

The power radiated by the relay transmitting antenna in a direction, determined by the angles θ and Φ, is equal to

$$P_{\text{вых р}} = P_{\text{вх р}} K_p G_2 F_2^2(\theta, \Phi). \qquad (8.10)$$

where K_p is the gain of the relay amplifier; G_2 is the maximum directive gain of the relay transmitting antenna; $F_2(\theta, \Phi)$ is a function describing the standardized antenna radiation pattern of the relay transmitting antenna with respect to field. Subsequently for simplicity it is assumed that

$$F_{p1}(\theta, \Phi) = F_{p2}(\theta, \Phi); \; G_2 = G_п.$$

Between the relay receiving and transmitting antennas it is necessary to provide an appropriate bypass, at least, not less than by K_p times with respect to power.

The powers of the interference and effective signals on the input of the receiver of the suppressed radar are respectively equal to

$$P_{\text{п вх}} = \frac{P_c G_c}{4\pi D_p^2} \frac{A_{r1} A_{r2}}{4\pi D_p^2} G_п F_c^2(\theta_p, \Phi_p) F_p^4(\theta, \Phi) K_p, \qquad (8.11)$$

$$P_{\text{с вх}} = \frac{P_c G_c}{4\pi D_c^2} \frac{\sigma_ц}{4\pi D_c^2} A_{r2} F_c^4(\theta'_c, \Phi'_c), \qquad (8.12)$$

where $\sigma_{\text{ц}}$ is the effective scatter area of the covered aircraft (Ц); D_{c}, θ'_{c}, Φ'_{c} are the polar coordinates of the covered aircraft Ц (the angles θ'_{c}, Φ'_{c} are calculated from the maximum antenna radiation pattern of the suppressed radar); A_{r2} is the maximum equivalent area of absorption of the antenna of the suppressed radar.

From (8.11) and (8.12) we have for the interference signal ratio on the input of the suppressed receiver

$$k = \left(\frac{P_{\text{п}}}{P_{\text{c}}}\right)_{\text{вх}} = \frac{A_{r1} G_{\text{п}} F_{\text{c}}^{2}(\theta_{\text{p}}, \Phi_{\text{p}}) F_{\text{p}}^{4}(\theta, \Phi) D_{\text{c}}^{4} K_{\text{p}}}{\sigma_{\text{ц}} F_{\text{c}}^{4}(\theta'_{\text{c}}, \Phi'_{\text{c}}) D_{\text{p}}^{4}}. \tag{8.13}$$

Hence we easily obtain the expression for the necessary value of the relay amplification factor

$$K_{\text{p}} = \frac{k_{\text{п}} \sigma_{\text{ц}} F_{\text{c}}^{4}(\theta_{\text{c}}, \Phi_{\text{c}}) D_{\text{p}}^{4}}{A_{r2} G_{\text{п}} F_{\text{c}}^{2}(\theta_{\text{p}}, \Phi_{\text{p}}) F_{\text{p}}^{4}(\theta_{\text{л}}, \Phi_{\text{л}}) D_{\text{c}}^{4}}. \tag{8.14}$$

In the particular case, when the distance between the trap and the covered aircraft is small as compared to the distance between the aircraft and the suppressed radar ($D_{\text{л}} \ll D_{\text{c}}$), expression (8.14) is simplified. Then

$$K_{\text{p}} = \frac{k_{\text{п}} \sigma_{\text{ц}}}{G_{\text{п}} A_{r1}} = \frac{k_{\text{п}} \sigma_{\text{ц}}}{G_{\text{п}}^{2}} \frac{4\pi}{\lambda^{2}}, \tag{8.15}$$

where $G_{\text{п}}$ is the directive gain of the relay receiving antenna.

Passive Reradiators

The possibility of increasing the effective scatter area of traps with the help of passive reradiators is based on the peculiarities of the scattering of an incident plane wave by conducting bodies. The ESA of any body for a given direction is determined by the known formula:

$$\sigma = s_{\text{э}} G, \tag{8.16}$$

where $s_2 = \frac{P_2}{p}$ is the ratio of the power (P_2) dispersed by a given body, to the flux density of the power (p) of electromagnetic energy incident on a reradiator; G is the directive gain of a reradiator in a given direction (in the direction to the point of observation).

For flat bodies and also for certain other bodies, close in their reradiating properties to flat bodies, value s_2 is equivalent to the absorption area A_r of a certain antenna

$$A_r = s_2 = \frac{G\lambda^2}{4\pi}. \qquad (8.17)$$

Substituting G from (8.17) into (8.16), we obtain

$$\sigma = \frac{4\pi}{\lambda^2} s_2^2 .$$

For an ideally conducting flat plate, the dimensions of which are considerably greater than the wavelength, in the case of its normal orientation in the direction of the incident wave the ESA is written in the form

$$\sigma_{\text{макс}} = \frac{4\pi s^2}{\lambda^2}, \qquad (8.18)$$

where s is the area of the plate.

In proportion to the change of orientation of the plate the magnitude of reflected energy rapidly varies. In Fig. 8.9 there is shown the reradiation pattern of a metallic plate, the dimensions of which are considerably greater than the wavelength. Due to the very acute reradiation pattern the metallic plate is unsuitable for increasing the ESA of aircrafts.

Reradiators mounted on aircrafts must meet the following requirements:

- have a large ESA with dimensions and weights as small as possible;

– possess a rather broad reradiation pattern.

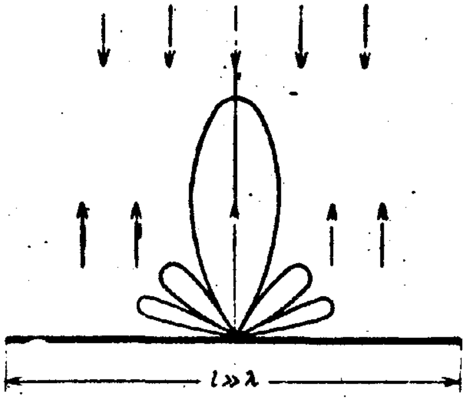


Fig. 8.9. The reradiation pattern of a metallic plate.

These requirements to a certain extent are met by so-called corner reflectors of various types, reradiators in the form of Luneberg lenses and responders.

Corner reflectors are of rigid construction consisting most often of three mutually perpendicular facets electrically connected with each other. Depending upon the shape of their facets triangular, rectangular and round corner reflectors (Fig. 8.10) are distinguished. Their maximum ESA are respectively equal to:

$$\sigma_{\Delta}=\frac{4}{3}\pi\frac{a^4}{\lambda^2}, \qquad (8.19)$$

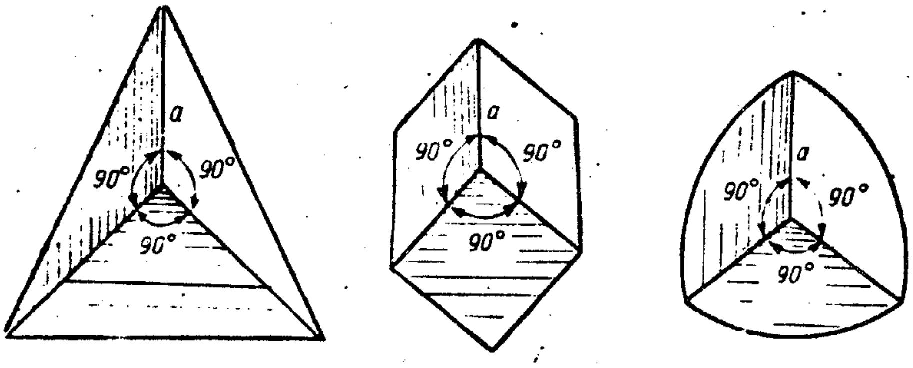
$$\sigma_{\square}=12\pi\frac{a^4}{\lambda^2}, \qquad (8.20)$$

$$\sigma_{\bigcirc}=2\pi\frac{a^4}{\lambda^2}. \qquad (8.21)$$

Here a is the length of a rib of a reflector.

Corner reflectors with small dimensions give a high ESA. Thus, when λ = 3 cm and a = 50 cm $\sigma_{\square}$ = 2500 m^2.

The width of the reradiation pattern of the corner reflectors at the level of half power is approximately 40-50°. In order to



a) b) c)

Fig. 8.10. Corner reflectors: a) triangular; b) rectangular; c) round.

increase the reradiation sector, several corner reflectors variously oriented in space are used. For example, the corner reflector depicted in Fig. 8.11 creates practically omnidirectional reradiation.

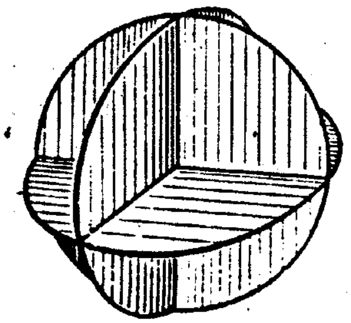
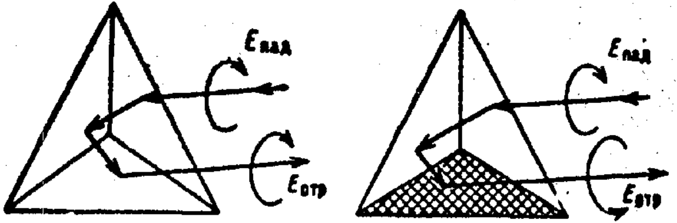


Fig. 8.11. An omnidirectional corner reflector.

The magnitude of the maximum ESA of corner reflectors essentially depends on how long the accuracy of the right angles between the facets of a reflector is maintained. An inaccuracy of an angle of as much as 1° leads to a decrease in the maximum value of the ESA of a corner reflector by 2-5 times.

The jamming (with corner reflectors) of radars, having antennas with circular polarization can be ineffective. This is explained by the fact that from the conducting facets of a corner reflector the wave is reflected an odd number, in consequence of which the direction of rotation of the vector of the electrical field of a reflected signal is reversed (Fig. 8.12a). The change of polarization of the reflected waves occurring on corner reflectors can be eliminated, if one of its facets is covered with a dielectric layer (Fig. 8.12b) [67].



a) b)

Fig. 8.12. Change of polarization of a radio wave reflected from a corner reflector: a) all facets are metallic; b) two facets are metallic; one facet is covered with a dielectric.

One of the significant deficiencies of corner reflectors is the small width of their reradiation pattern at a level of half power. Reradiators realized on the basis of Luneberg lenses possess a rather broad reradiation pattern.

A Luneberg lens is a dielectric sphere. The refractive index of the dielectric (n) in an ideal Luneberg lens depends only on the ratio of the current radius of the lens (r) to the external radius of the lens (R)

$$n=\sqrt{2-\left(\frac{r}{R}\right)^2}. \tag{8.22}$$

In Fig. 8.13a there are depicted trajectories of rays in a Luneberg lens. The central ray AO does not experience refractions in the lens, whereas the trajectories of the remaining rays are distorted. As a result all the rays are focused at point O on the interior side of the sphere covered with a metallic film. Point O being the source of the secondary electromagnetic waves will create on the output of the lens a cophasal distribution of the field, so

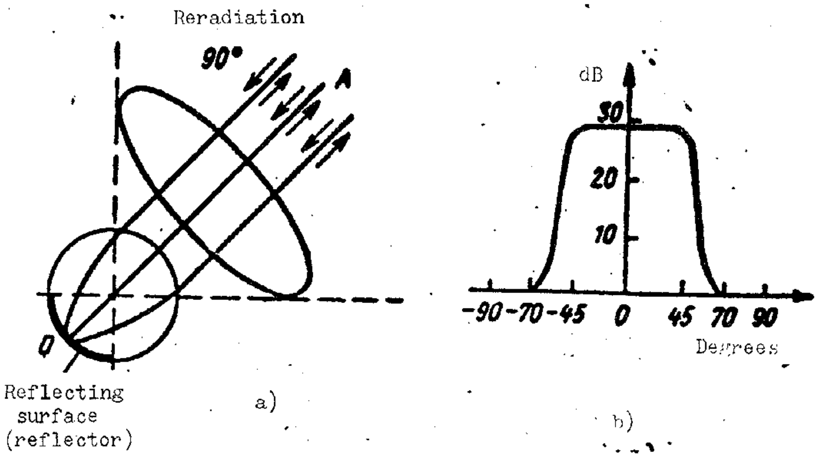


Fig. 8.13. The trajectory of rays (a) in a Luneberg lens with a 90-degree reflector and its reradiation pattern (b).

that maximum reradiation pattern will coincide with the incoming direction of the incident wave.

The maximum effective scatter area of a Luneberg lens can be found by substituting $s = \pi R^2$ in (8.18), i.e.,

$$\sigma_{\text{л}} = 4\pi^3 \frac{R^4}{\lambda^2}.$$

The width of the reradiation pattern of a Luneberg lens depends on the dimensions of the shielding (metallic) surface of the sphere. Thus, for a shielding surface of the size of a quarter of the surface of the sphere the width of the reradiation pattern at the level of half power is about 90° (Fig. 8.13b). In Fig. 8.14b there is represented the reradiation pattern of a Luneberg lens with a 140-degree reflector [68]. The reradiation sector of this lens is approximately equal to 140°.

A Luneberg lens does not provide reradiation in a circular pattern. The better can be attained using this lens, if part of its sphere is encircled with a metallic ring [69].

In Fig. 8.15a there is depicted an omnidirectional (in an azimuthal plane) Luneberg lens with a reflecting ring centered with respect to

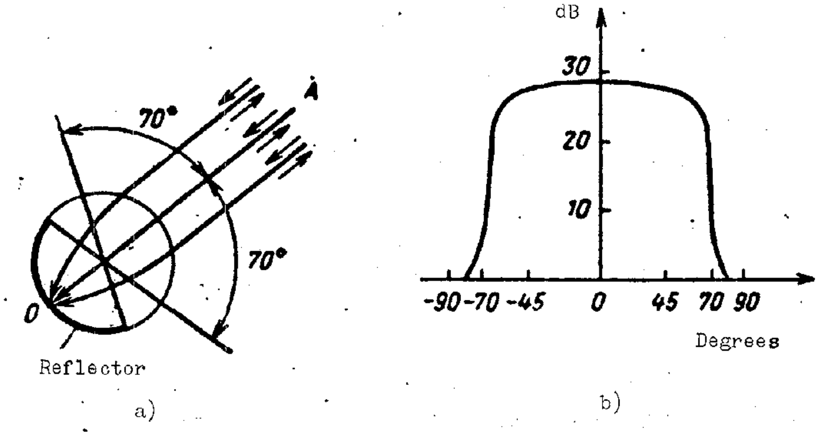


Fig. 8.14. Trajectories of rays (a) in a Luneberg lens with a 140-degree reflector and their reradiation pattern (b).

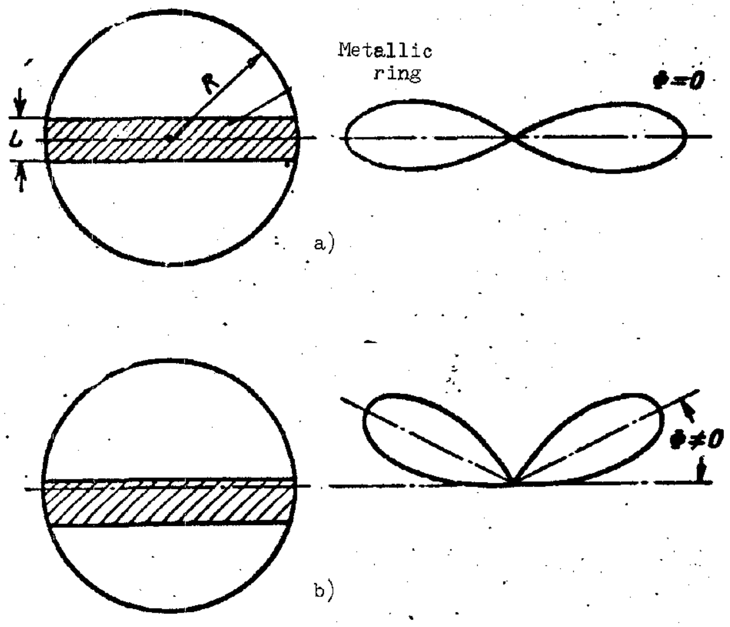


Fig. 8.15. An omnidirectional (in a azimuthal plane) Luneberg lens with a metallic ring centered (a) and displaced (b) with respect to the equator.

the equator. The position of the metallic ring determines the direction of the maximum reradiation. Thus, for a ring centered with respect to the equator the maximum reradiation will be situated in the equatorial plane (Fig. 8.15a). If, however, the ring is displaced, the lobe of the reradiation pattern is deflected from the equatorial plane (Fig. 8.15b). The maximum value of ESA is determined by the formula

$$\sigma_{\text{к}} = 4\pi \frac{(\pi R^2 - 2RL)^2}{\lambda^2}, \qquad (8.23)$$

where R is the radius of the sphere; L is the width of the metallic ring.

A lens in the form of a dielectric sphere with a metallic ring has somewhat lesser effectiveness than the earlier examined Luneberg lens with a reflector. A comparison of (8.23) and (8.22) gives

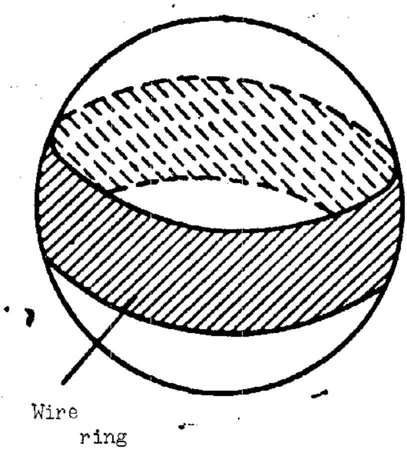
$$\eta = \frac{\sigma_{\text{к}}}{\sigma_{\text{л}}} = \left(1 - \frac{2}{\pi} \frac{L}{R}\right).$$

For example, if L/R = 0.2, then $\eta \approx 0{,}9$. Consequently, the decrease of the ESA of $\sigma_{\text{к}}$ as compared to a normal lens is insignificant.

Increasing the width of the metallic ring leads to an expansion of the reradiation pattern, but it also simultaneously decreases the ESA of the lens. This contradiction is well resolved in a Luneberg lens with a ring in the form of a network of parallel wires wound at 45° angle (Fig. 8.16 and 8.17). Such a lens is sometimes called a helisphere.

An electromagnetic wave with linear polarization falling on a helisphere at a 45° angle passes through the frontal part of the ring and is reflected from the opposite part of the ring. For other polarizations (vertical, horizontal, circular) there will be polarization losses, the maximum of which (during double passage) is 6 dB.

To obtain an isotropic pattern helispheres with two orthogonal wire networks are used. By selecting the width (and sometimes the configuration) of the rings it is possible to obtain relatively small heterogeneity in the reradiation pattern.



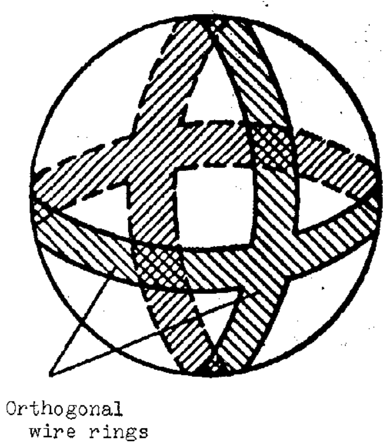


Fig. 8.16. Fig. 8.17.

Fig. 8.16. A helispheric Luneberg lens with a ring in the form of a network, composed of parallel wires.

Fig. 8.17. A helispheric omnidirectional Luneberg lens with orthogonal wire networks.

In principle the creation of an ideal lens without metallic rings with isotropic reradiation (Iton-Lipman lens) is possible. For this the coefficient of the dielectric filler should vary according to the principle

$$n=\sqrt{\frac{2R}{r}-1},$$

where r is the current radius.

Trajectories of rays in lens with circular reradiation are depicted in Fig. 8.18.

Dielectric helispheric lens are rather heavy. They are difficult to realize due to the necessity of putting in the central part $(r\approx 0)$ a dielectric with a very large coefficient of refraction.

The so-called hollow helispheric reflectors (Fig. 8.19a) are considerably lighter in weight. The reflecting ring is made out of metal and is a spherical segment. The wire network and metallic ring

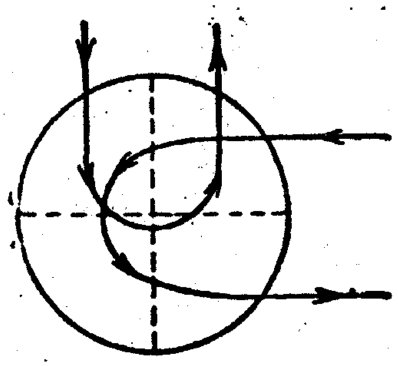


Fig. 8.18. The behavior of beams in an Iton-Lipman lens (with circular reradiation).

are situated orthogonally. The ESA of such a reflector is 10 dB less than the ESA of an ideal Luneberg lens. The course of beams of incident and reflected waves is shown in Fig. 8.19b.

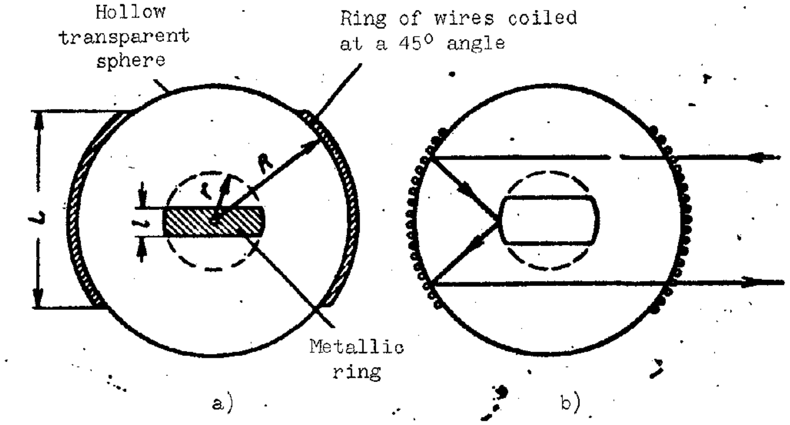

Hollow transparent sphere
Ring of wires coiled at a 45° angle
R
r
L
l
Metallic ring
a)
b)

Fig. 8.19. Hollow helispheric lens: a) construction; b) behavior of rays.

In Fig. 8.20 there are given the dependences on D/λ of the ESA of various reflectors standardized relative to the ESA of a metallic sphere of the same diameter D. Along the axis of the ordinates there is plotted the value

$$a = 10 \lg \frac{\sigma}{\sigma_m},$$

where σ_m is the ESA of the metallic sphere, σ is the ESA of the lens.

The curves represent: 1 – an ideal Luneberg lens, 2 – a helisphere with a ring inside (Fig. 8.19), 3 – a hollow helisphere (linear

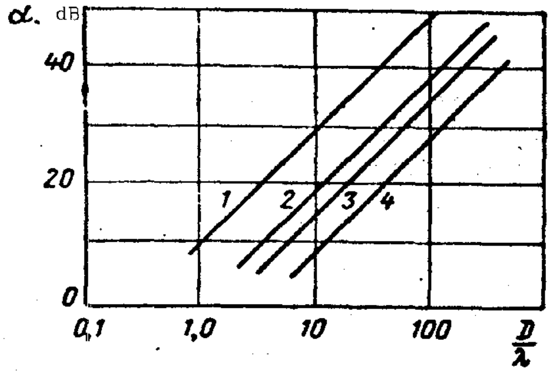


Fig. 8.20. The dependences of the ESA of various reflectors on their dimensions: 1 – ideal Luneberg lens; 2 – helisphere with ring inside; 3 – hollow helispheric lens (for linear polarization); 4 – hollow helispheric lens (for circular polarization); $\alpha = 10 \lg \sigma/\sigma_{ш}$.

polarization), 4 – hollow helisphere (circular polarization).

A Van Atta responder is, in point of fact, an antenna array, made out of a large number of dipoles or spirals (Fig. 8.21). The dipoles are located at an equal removal from the axis of symmetry of the responder and are connected in pairs by a coaxial cable of identical length. An electromagnetic wave received by dipole 1, is reradiated by dipole 6. In turn, dipole 1 reradiates a wave received by dipole 6. The electrical lengths of the 1-6 antenna-feeder, just like the other dipoles connected in paris are identical. The signals received and reradiated by the dipoles follow-on identical path. Therefore the direction of the maximum reradiation pattern will coincide with the incoming direction of the incident wave.

The arrays are designed for the reflection of waves with any polarization. For this the dipoles are arranged on a metallic screen at various angles (as a rule, each pair is at an angle of 90° to its neighboring dipole).

The effective scatter area of a responder made-up of n half-wave dipoles located at a distance of λ/2 from each other and at a distance

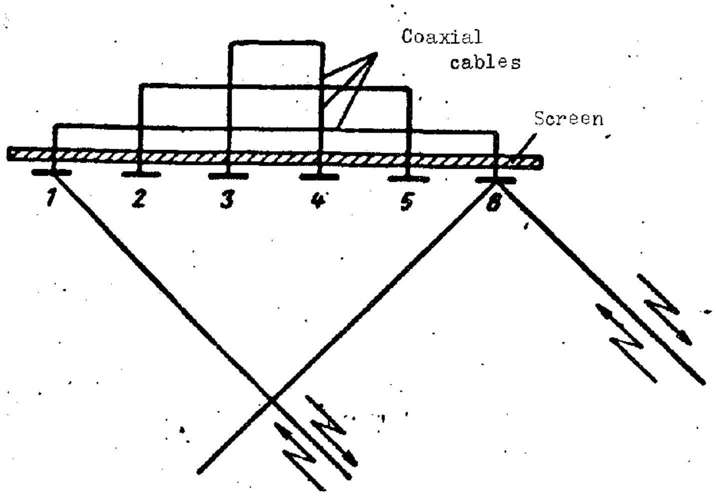


Fig. 8.21. A passive Van Atta responder.

of λ/4 from the reflecting screen can be found by the formula [70]

$$\sigma = 4\pi \frac{S^2}{\lambda^2}\left[\sin\left(\frac{\pi}{2}\cos\theta\right)\right]^4,$$

where θ is the angle of incidence; S is the area of aperture of the array.

Assuming that $S \approx n\lambda^2/4$, we obtain an expression for the maximum ESA of a Van Atta array

$$\sigma \approx \frac{\pi}{4} n^2\lambda^2.$$

A reradiated signal can be modulated in amplitude. For this phase shifters are included in the feeder lines connecting the dipoles. By appropriate variation of phase shift it is possible to obtain the necessary amplitude modulation of a reradiated signal.

A Van Atta reradiator can also be made in an active variant (relays), when the received signal is amplified in each channel and reradiated (Fig. 8.22). The main difficulty in making this type of reradiator consists in bypassing the receiving and transmitting pathways.

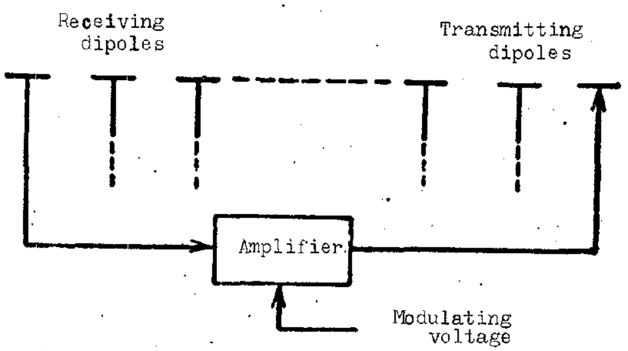


Fig. 8.22. An active Van Atta responder.

A deficiency of the Van Atta responder is its relatively short range.

In conclusion let us note that the various reradiators examined in this section do not exhaust the great diversity of means and methods of increasing the ESA of false targets. For example work is being conducted investigating the possibility of increasing the ESA of aircraft by the ionization of the space around a false target, in particular by the ionization of the stream of the jet engine of a rocket-trap by the addition (injection) of readily ionizing elements into the composition of the fuel [71].

CHAPTER 9

METHODS OF JAMMING BASED ON CHANGING THE ELECTRICAL PROPERTIES OF A MEDIUM AND RADAR CAPABILITY TO OBSERVE A TARGET

9.1. Counteracting the Operation of Electronic Equipment by Ionizing Local Regions of Space

At present two basic methods of ionizing space are known, which are applied to foil electronic systems [71]:

- the atomization and burning of readily ionizable elements (cesium, sodium, etc.),
- high-altitude nuclear explosions.

The physical principles of jamming by artificial ionization of space are based on the phenomena of absorption, reflection and refraction of electromagnetic waves in plasma.

Let us first dwell on the phenomena of refraction and reflection. As is known from electrodynamics the reflection of electromagnetic waves occurs in all cases, when the macroscopic parameters of heterogeneity $n = \sqrt{\varepsilon\mu}$ and σ are different from the corresponding parameters of the medium, in which the radio waves are propogated. Radio wave refraction also occurs as a result of this same principle. Thus, to provide considerable reflection of radio waves by ionized formations significant local change in the macroscopic parameters of the medium n and σ is necessary. The application of the indicated macroscopic parameters as electrical characteristics of the medium is permissible, if the average distance between the particles forming the medium is much less than the wavelength (d << λ), i.e., when for the incident and propagating waves the medium is continuous.

Errors arise in determining the direction to the source of the radio waves in the case when the local heterogeneity in its turn is itself heterogenous, i.e., its refractive index is a function of the coordinates (Fig. 9.1.).

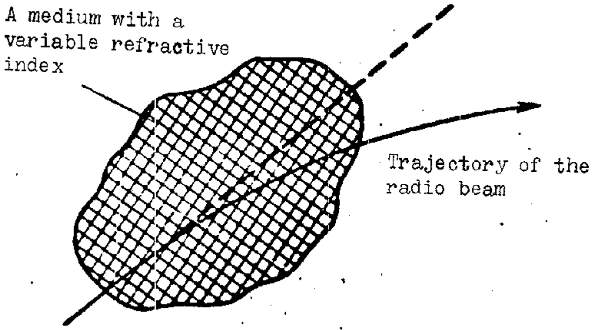

A medium with a variable refractive index
Trajectory of the radio beam

Fig. 9.1. The behavior of beams in local heterogeneity.

Errors in determining direction can theoretically occur and with the constancy of the electrical parameters of local heterogeneity, if this heterogeneity has the corresponding form (for example, nonrectangular).

The refractive index of an ionized medium (local heterogeneity) neglecting the influence of the magnetic field of earth is approximately determined by the following formula:

$$n \approx \sqrt{1 - 81 \frac{N}{f^2}}.$$

Here f is the carrier frequency in Hz; N is the number of electrons in one cubic meter.

With a sufficiently high concentration of electrons the radio waves can be completely reflected from the ionized region (total internal reflection). The critical frequency corresponding the total reflection of radio waves is determined from the condition n = 0, it follows from this that

$$f_{\text{кр}} = 9\sqrt{N}.$$

Thus, to obtain total reflection of the ocsillations with the carrier frequency f from an ionized region it is necessary to have this concentration of electrons

$$N = \frac{f_{\text{кр}}^2}{81}.$$

For example, for the wave $\lambda = 3$ cm we obtain $N = 10^{18}$ e/m^3.

In order to provide a certain concentration of electrons (N), it is necessary to have powerful sources of ionization.[1]

One may assume with sufficient accuracy that the necessary ionization source strength is equal to

$$I = \alpha N^2,$$

where α is the electrons recombination coefficient equal at the surface of the earth to approximately $\alpha = 10^{-12}$ cm^3/s·e.

When $N = 10^{18}$ e/cm^3 and $\alpha = 10^{-12}$ cm^3/s·e the ionization source strength must be $I = 10^{24}$ e/m^3·s or $I = 10^{18}$ e/cm^3·s.

In other words, to create an ionized region with a concentration of 10^{18} electrons per cubic meter the ionization source should in one second create 10^{24} electrons per cubic meter. Such a high concentration of electrons can be temporarily created by nuclear expolsions or by the simultaneous combustion of large quantities of readily ionizable element, for example cesium and others.

With the explosion of nuclear ammunition there will form at the epicenter of the explosion high concentrations of electrons. However due to their recombination the electron concentration rapidly drops in time, so that the interference of a nuclear explosion on radar in the centimeter range is very short-term. There is a considerably greater effect on the propagation of waves in the meter range by high-altitude nuclear explosions.

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[1] Let us note that the threshold concentration of electrons necessary for the visual detection of ionization, is equal $10^{17} \div 10^{18}$ electrons on cubic meter.

Let us dwell on the absorbing properties of locally ionized media.

The mechanism of radio wave absorption in an ionized region can be clarified in the following manner. Free electrons under the effect of an electrical field of an incident wave caused forced oscillations with a frequency equal to the carrier frequency of the electromagnetic oscillations. During the course of their oscillatory motions electrons collide with neutral molecules, atoms and ions and increase their kinetic energy. In this way the transition of the electromagnetic field energy into the thermal energy of the medium is carried out.

The absorbent property of the ionized region is characterized by the coefficient of absorption of radio waves (dB/km):

$$\beta=\frac{1,8\cdot 10^{-2}N\nu}{\omega^2+\nu^2}, \qquad (9.1)$$

where N is the number of electrons per m^3; ν is the number of collisions of electrons with other particles (ions, atoms and molecules of gas) per second; $\omega = 2\pi f$ is the angular frequency.

From formula (9.1) it follows that the coefficient of absorption has its maximum at a certain value of collision frequency. Using the rule of finding the extremum of a function, we obtain that

$$\beta=\beta_{\text{макс}} \text{ or } \omega=\nu.$$

Collision frequency ν is proportional to air density. Therefore there is a certain interval of atmospheric altitudes, within the limits of which the damping of radio waves is the greatest. The calculations and experimental investigations show that the damping of radio waves takes place in its greatest intent within the limits of the 16-kilometer band with the center located approximately at a height of 72 km [72]. Collision frequency at a height of 72 km is approximately equal to $\nu \approx 6 \cdot 10^6$ collisions per second.[1]

For signals with carrier frequencies of $f > 5$ MHz the value of ω^2 in formula (9.1) is considerably greater than ν^2, due to which

$$\beta = \frac{0{,}45 \cdot 10^{-3} N}{f^2}. \qquad (9.2)$$

With the help of (9.2) it is possible to determine the necessary concentration of electrons, providing a certain damping β in dB/km:

$$N = \frac{\beta f^2}{0{,}45} 10^3 \approx 2{,}2 \beta f^2 10^3 \; [e/m^3].$$

To obtain, for example, the damping $\beta = 10$ dB/km at a height of 72 km for a wave $\lambda = 3$ cm it is necessary to create an ionized region with a concentration of electrons of $N = 0.37 \cdot 10^{18}$ e/m^3. Such a high concentration for large dimensions at the present time can be created only for a very short time with the help of powerful nuclear explosions.

In practice noticeable absorption of radio waves can occur for meter and longer waves. The coefficient of absorption of longer waves attains a considerable magnitude with comparatively low concentrations of electrons ($N = 10^{11}$-10^{14} e/m^3).

9.2. The Influence of Nuclear Explosions on the Operation of Electronic Systems

Explosions at altitudes lower than 16 km do not cause long-term ionization, therefore they cannot have a significant effect on the operation of electronic systems. With ground-level and underground (underwater) explosions regions can form, in which intense absorption and reflection of radio waves occur. However, the effect of the absorption and reflection of radio waves is not connected with the ionization of space, but with presence of local heterogeneity in the medium having a high concentration of particles of solid substances and water projected into the atmosphere. The effect on the operation of electronic systems in the centimeter range due to the products of an explosion can be shown only for the first stage of the explosion.

---

[1] At low altitudes damping of the shorter waves does not increase inasmuch as the lifetime of electrons decreases to a significant extent and the initial assumptions on which formula (9.1) is based are incorrect.

The duration of the existence of ionized regions depends on the altitude and the power of the explosion, the time of day, etc.

At high altitudes (more than 40-50 km) rather stable regions will be formed with a relatively high concentration of electrons.

With the explosion of nuclear material at high altitudes ionized regions with small concentration of electrons (10^{10}-10^{11}) can exist for several minutes and even hours.

In the first approximation it is possible to distinguish two types of regions.

First, regions of slow electrons formed due to the ionization of the medium, mainly, by thermal X-rays. These regions have comparatively limited dimensions, tens and hundreds of km, and the electron concentration in them decreases approximately according to the principle

$$N = 10^{13} \frac{1}{t} \, [e/m^3].$$

Here t is time in seconds.

The development of a region in time after it was formed, occurs mainly according to the laws of diffusion.

Secondly, regions of fast (relativistic) electrons (β-particles), radiated by the radioactive products of fission. The fast electrons are captured by the magnetic field of the earth, in connection with which the ionization of space on an earth-wide scale takes on a global character (and not local, as in the preceding case). Let us dwell on this question in somewhat greater detail. Let us recall certain initial physical premises.

On a charge of e moving with a speed of $\overline{v}$ in a magnetic field with a strength of $\overline{H}$, there acts a Lorenz force F determined by the formula

$$F = e[\bar{v}\ H].$$

This force other things being equal is proportional to the velocity of a particle. If the magnetic field is uniform and its lines of force are parallel each other, then the trajectory of a charged particle entering into the magnetic field with a speed of $\overline{v}$, in general will be a cylindrical spiral with a constant pitch, wound around a line of force. The radius of the cylinder (Larmor radius) is determined by the formula

$$r = \frac{mv}{eH}.$$

In a nonuniform field the trajectory of travel of a charged particle is quite complicated. In particular, with the travel of a charged particle in a nonuniform magnetic field in the direction of increase of field strength (in the direction of travel of a particle the lines of force of a magnetic field converge) on a charged particle there will act a force tending to push it into the region of smaller magnitudes of field strength [78]. The appearance of this force is caused by a component of the magnetic field H_2 (Fig. 9.2), the magnitude of which is the greater the greater is the heterogeneity of the field (the greater is angle γ). If field H were uniform, then all its lines of force would be parallel to component $H_1 (\gamma = 0)$ and component H_2 would be equal to zero. The heterogeneity of the field generates component H_2.

Let us examine the simplest case – the movement of an electron is a circular orbit in a heterogeneous field (Fig. 9.2). Component H_1 of magnetic field H generates a centripetal force providing the rotary motion of an electron. Component H_2, acting at point 0 on an electron moving with speed v, generates a force $F(H_2)$ ejecting the

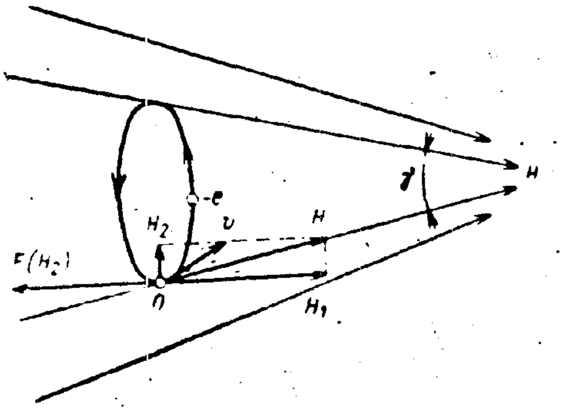


Fig. 9.2. Forces acting on a charged particle in a nonuniform magnetic field.

particle into the region of the field with smaller strength. The presence of an ejecting force in a nonuniform magnetic field and the heterogeneity itself lead to an essential transformation of the spiral trajectory.

In Fig. 9.3 there is depicted an approximate trajectory of an electron in a nonuniform magnetic field. In proportion to the advance of an electron into the region of greater concentration of magnetic lines of force the pitch of the spiral and the Larmor radius decrease.

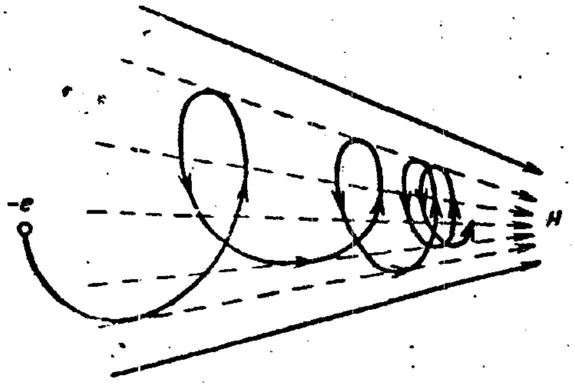


Fig. 9.3. Travel of an electron in a nonuniform magnetic field.

Inasmuch as the Lorentz forces in the first approximation do not change the absolute value of the velocity of a particle, then in proportion to the deceleration of the longitudinal travel of an electron due to an inhibitory force, the component of speed v_1, perpendicular to H_1, will certainly increase. More thorough investigations show that ratio v_1^2/H can be considered approximately constant during the whole time of travel of a particle. This circumstance makes it possible to find angle α between the velocity vector of a particle and the direction of a line of force at a given moment of time, if the initial angle α_0 is known at the time of the entry of a particle into a magnetic field.

Accordingly we have (Fig. 9.4)

$$\frac{\sin^2\alpha_0}{H_0}=\frac{\sin^2\alpha}{H}. \tag{9.3}$$

Hence

$$\sin\alpha=\sin\alpha_0\sqrt{\frac{H}{H_0}}.$$

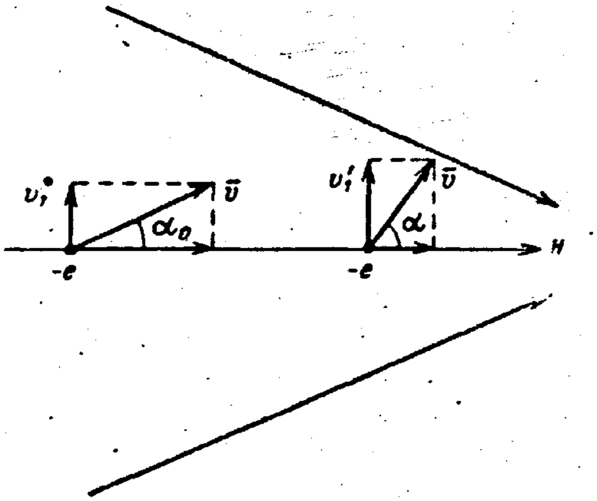


Fig. 9.4. Travel of an electron in a nonuniform magnetic field.

For a certain angle α_0 we will determine the magnetic field strength, at which the forward travel into the region of greater concentration of magnetic lines of force ($\sin\alpha = 1$) will stop:

$$H=\frac{H_0}{\sin^2\alpha_0}. \tag{9.4}$$

An electron cannot penetrate into a region of greater magnetic field strength. Having reached the indicated region of the magnetic field, an electron starts to travel in the opposite direction. Thus, a region of high concentration of magnetic lines of force can play the role of a unique magnetic mirror.

The magnetic field of the earth has two regions of high concentration of magnetic lines of force — the northern and southern magnetic poles (Figs. 9.5 and 9.6).

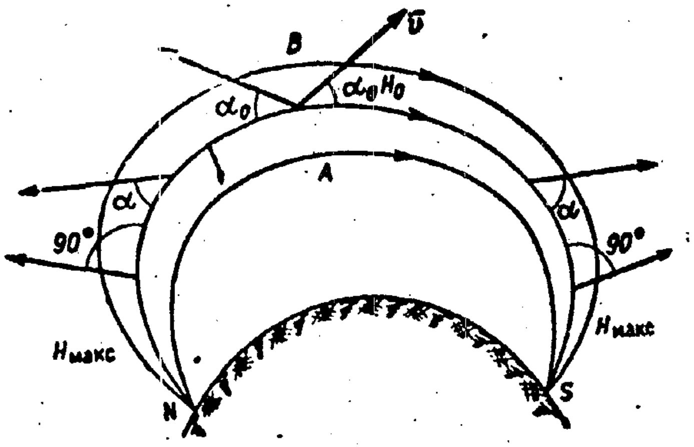


Fig. 9.5. Travel of a free electron in the magnetic field of the earth.

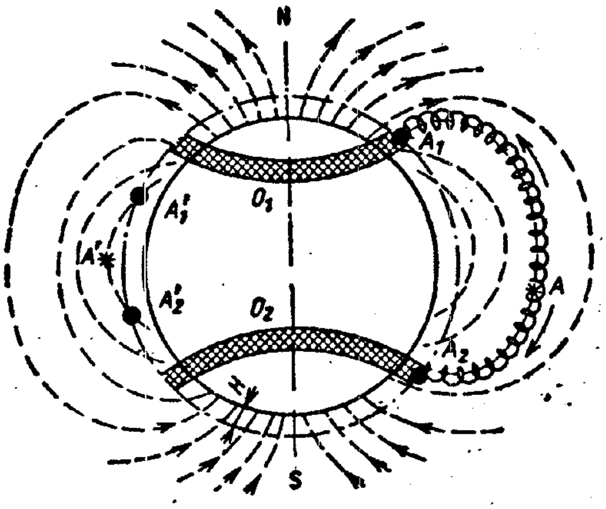


Fig. 9.6. The formation of ionized regions in the magnetic field of the earth.

If the maximum magnetic field strengths near the poles are identical ($H_{макс}$), then the possibility of the reflection of an electron from a magnetic mirror will be determined by angle α_0. All particles, for which $\sin\alpha_0 > \sqrt{\frac{H_0}{H_{макс}}}$, will be reflected from the magnetic mirrors forming near the poles.

Thus, the magnetic field of the earth, for the indicated particles, will actually be a "magnetic trap." The regions of space, in which the reflection of electrons "from the magnetic mirrors" occurs, are called conjugate points (A_1 and A_2 in Fig. 9.6). Conjugate points in principle cannot coincide with the magnetic poles. Formula (9.3) makes it possible to find conjugate points in other sections of space near the earth.

Electrons captured by the magnetic field of the earth shifting along the magnetic lines of force will simultaneously experience "magnetic drift" from east to west. The "magnetic drift" of electrons is caused by a decrease in magnetic field strength with height.

The physical cause of the "magnetic drift" of electrons due to the decrease of the field in radius is illustrated by Fig. 9.7. In Fig. 9.7 there is depicted a magnetic field (diminishing in radius) of direct current I flowing along a straight wire. If an electron (or positive ion) rotates near the lines of force A and B, then the Larmor radii of its trajectory for sections A and B will be different. At section A the Larmor radius will be less than at section B with a smaller magnetic field strength. After completion of the first orbit an electron moving as is shown in Fig. 9.7 will be somewhat lower than initial point of travel. With subsequent turns it will be continuously displaced in a direction opposite the direction of current I generating this magnetic field. Hence it directly follows that under terrestrial conditions electrons moving along lines of force will "drift" from east to west. Accordingly positively charged particles will "drift" from west to east. An analogous phenomenon occurs in the so-called radiation belts of the earth.

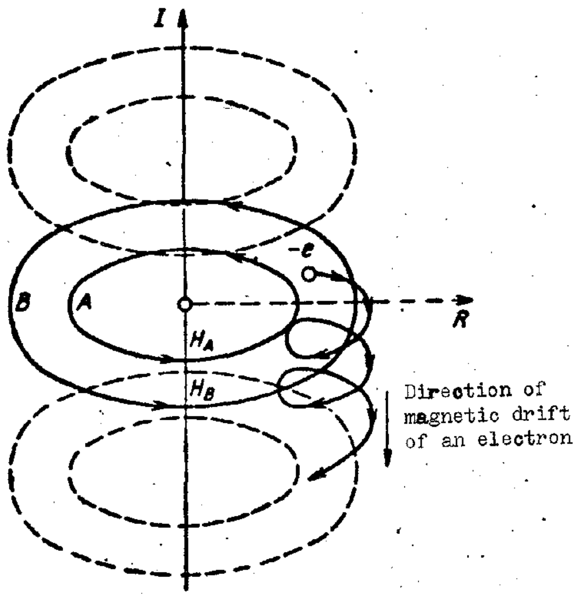


Fig. 9.7. "Magnetic drift" of an electron.

Thus, due to fast electrons forming in a high-altitude nuclear explosion, the ionization of space near the earth takes on a global character. However the density of the electrons on the average is small, and the described effects have no noticeable influence on the operation of radar systems, with the exception, perhaps, of radar in the meter range, operating in the region of the formation of the conjugate points. Concerning means of communication and radio navigation, especially in the short-wave and medium-wave ranges, there can arise here serious disturbances of operation for a comparatively prolonged period of time.

The lifetime of global ionized regions formed by fast electrons depends on the height of the conjugate points above the earth. If the conjugate points are located in a region of increased atmospheric density, then a decrease of electrons occurs more rapidly due to absorption by neutral molecules and positive ions located in the region of the magnetic mirrors. The absorption of fast electrons leads to the ionization of space in the vicinity of the conjugate points. The decrease of electrons beyond the conjugate points, as a rule, is insignificant. After the detonation of nuclear material at a height of 480 km (operation "Argus") there was observed the formation of sharply expressed layers with a thickness of about 100 km. The ionized regions were preserved for several days [79].

9.3. Methods of Decreasing the Effective Scatter Area of Aircraft

Decreasing the ESA of aircraft is one of the important directions for combatting the electronic systems of an enemy. It is not so much due to the possibility of decreasing the detection range, as due to the possibility of proportionally decreasing the necessary power potentials of jamming transmitters, the necessary number of dipole reflectors, the necessary ESA of a trap, etc. This is caused by the fact that the detection range decreases proportionally to $\sqrt[4]{\sigma_{\text{ц}}}$, whereas the necessary power potential of a jamming transmitter, the necessary number of dipole reflectors and so forth decrease directly proportionally to $\sigma_{\text{ц}}$.

There are three methods of decreasing the ESA of aircraft:

- the selection of the form of the aircraft;
- the application of antiradar coverings;
- control by radio wave scattering.

Theoretically and practically it has been established that a sharp decrease in radio wave scattering is characteristic for bodies having small dimensions, small radii of surface curvature and not having sharp surface discontinuities.

It has been established in practice that the better the aerodynamic shape of the aricraft, the less its ESA will be. However, with respect to contemporary aircraft, in spite of their good aerodynamic shape, their ESA nevertheless remains rather high. Further decrease of ESA is achieved by application of antiradar coverings.

Two forms of antiradar coverings exists: absorbing and interference.

The material of absorbing coverings is selected because it provides complete absorption of waves incident on it and due to the absence of reflection of the latter from the boundary of the media.

In interference coverings the material and structure of the covering are selected in such a manner so that the incident and reflected waves mutually compensate one another.

Absorbing Coverings

Let us examine the reflection of electromagnetic wave from an infinite ideally conducting surface covered with a substance characterized by complex dielectric strength (ε') and magnetic (μ') permeability (Fig. 9.8):

$$\varepsilon' = \varepsilon'_r + j\varepsilon'_{к}, \tag{9.5}$$

$$\mu' = \mu'_r + j\mu'_{к}. \tag{9.6}$$

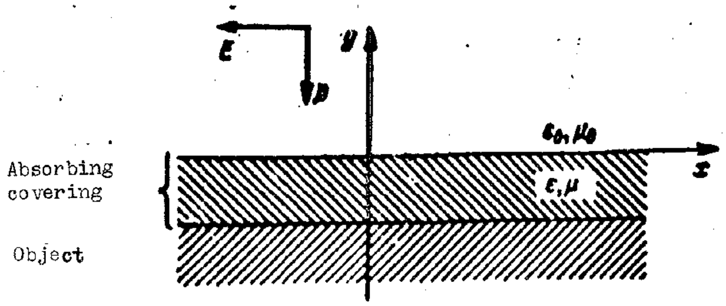


Fig. 9.8. Action of an absorbing covering.

Here $\varepsilon' = \varepsilon_0\varepsilon$ – is the dielectric strength of a covering (in free space $\varepsilon' = \varepsilon_0$); $\frac{\varepsilon'_r}{\varepsilon_0}=\varepsilon_r$ is the relative dielectric strength of the covering; $\frac{\varepsilon'_K}{\varepsilon_0}=\varepsilon_K$ is the imaginary part of the dielectric constant caused by dielectric losses and the electrical conductivity of the covering; $\mu' = \mu_0\mu$ is the permeability of the covering (in free space $\mu' = \mu_0$); $\frac{\mu'_r}{\mu_0}=\mu_r$ is the relative permeability of the covering; $\frac{\mu'_K}{\mu_0}=\mu_K$ is the imaginary part of permeability caused by losses.

Let us deterime the value of the parameters μ' and ε' of an absorbing covering, where the reflectivity from the boundary ($y = 0$) is equal to zero.

Let us write the expression for the complex reflectivity of a plane wave from the flat boundary of two media

$$\bar{R}=\frac{\bar{z}-z_0}{\bar{z}+z_0}, \tag{9.7}$$

where z_0 is the characteristic impedance of a free space:

$$z_0=\sqrt{\frac{\mu_0}{\varepsilon_0}}=120\pi; \tag{9.8}$$

$\bar{z}$ is the characteristic impedance of an absorbing covering and

$$\bar{z}=\sqrt{\frac{\mu'}{\varepsilon'}}. \tag{9.9}$$

Substituting (9.8) and (9.9) in (9.7), we obtain

$$R=\frac{1-\sqrt{\frac{\varepsilon'\mu_0}{\mu'\varepsilon_0}}}{1+\sqrt{\frac{\varepsilon'\mu_0}{\mu'\varepsilon_0}}}. \tag{9.10}$$

If one assumes that

$$\sqrt{\varepsilon\mu}=n+j\kappa,$$

where n is the refractive index, K is the attenuation factor of the medium, then expression (9.10) can be written in the form

$$\bar{R}=\frac{\mu-n-j\kappa}{\mu+n+j\kappa}. \tag{9.11}$$

From (9.11) it follows that reflectivity becomes zero, if this conditions is fulfilled

$$\mu=n+j\kappa \tag{9.12}$$

or taking into account (9.6)

$$\begin{aligned}\mu_r&=n,\\ \mu_\kappa&=\kappa.\end{aligned} \tag{9.13}$$

Thus, the relationships (9.12) and (9.13) are the conditions of complete absorption of an incident wave by a covering. These conditions are satisfied by coverings, the composition of which includes ferromagnetic materials and substances with rather large losses. Usually these are magnetodielectrics constituting a conglomerate of ferromagnetic material, the particles of which are isolated from each other by insulating material of a nonmagnetic dielectric. Single-layer coverings made from such materials are effective for waves in the meter and decimeter ranges.

For the absorption of waves in the centimeter range multilayer coverings are used with the parameters variable from layer to layer. Each layer of these coverings is made of expandable polystyrene, and the absorber is made of graphite or carbon black, the concentration of which from layer to layer varies.

To coordinate the covering with the exterior (free) space the relative dielectric constant of the external layer should be equal to 1 ($\varepsilon'_r = \varepsilon_0$), and the imaginary component (loss tangent) is close to zero. The relative dielectric constants and the loss tangents of the subsequent layers have to increase from layer to layer. Sharp change in the parameters ε and μ from layer to layer is impermissible, since this leads to an increase in the reflectivity of the radio waves from the boundary of the two media.

For the purpose of increasing the area of "contact" of an antiradar covering with an incident electromagnetic wave are widespread coverings with so-called "geometric heterogeneities." These coverings are characterized by the fact that their structure represents periodically repeated irregularities in the form of pyramids or cones (Fig. 9.9).

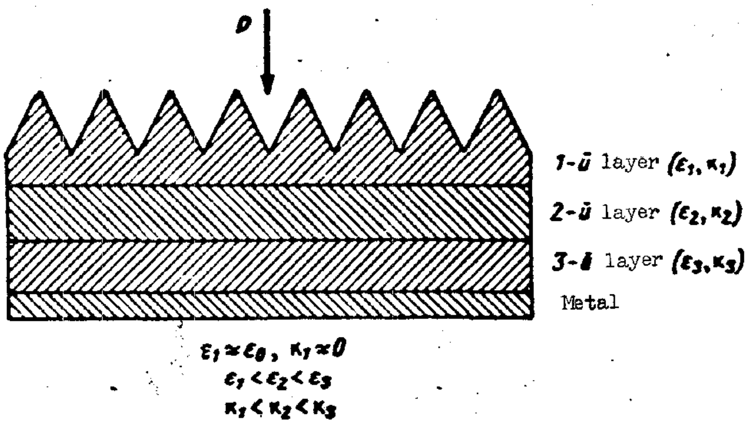

D
1-й layer (ε_1, κ_1)
2-й layer (ε_2, κ_2)
3-й layer (ε_3, κ_3)
Metal
$\varepsilon_1 \approx \varepsilon_0$, $\kappa_1 \approx 0$
$\varepsilon_1 < \varepsilon_2 < \varepsilon_3$
$\kappa_1 < \kappa_2 < \kappa_3$

Fig. 9.9. An absorbent covering with "geometric heterogeneities."

Well-known is the English two-layered absorbing covering of the AF type [63], made from a mixture of porous rubber and coal dust (carbon black). The reflectivity of such a covering with normal wave incidence in the $\lambda = 3\text{-}10$ cm range is a total of 6%. A corrugated two-layered covering of the same type has a reflectivity of about 1% in a rather large sector of incidence angles.

In Fig. 9.10 there is given the dependence of the reflectivity on wavelength for a covering of the AF-20 type. This corrugated covering is made on a base of grains of polystyrene pressed with coal dust. The thickness of the absorbing coverings attains several centimeters.

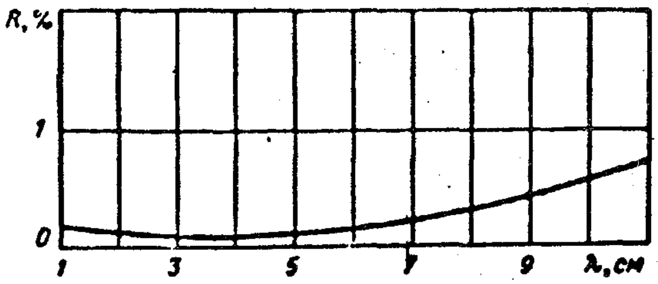


Fig. 9.10. The dependence of reflectivity on wavelength for an AF-20 absorbing covering.

Interference Coverings

In interference coverings an effect of lowering of the ESA of a protected object is attained due to the mutual attenuation of waves, reflected from the surface of the object and from the surface of the covering (interference of the incident and reflected radio waves). The incident wave is multiply reflected from the "covering-object" boundary of the two media and partially absorbed in the substance of the covering (Fig. 9.11). Let us determine the parameters of a covering, in which the total field in the direction to the source of the incident wave is equal to zero, i.e.

$$E_{\text{отр}}=\sum_{i=1}^{n}E_i=0, \tag{9.14}$$

where E_i is the component of the wave reflected from the "free space-covering" boundary (Fig. 9.11).

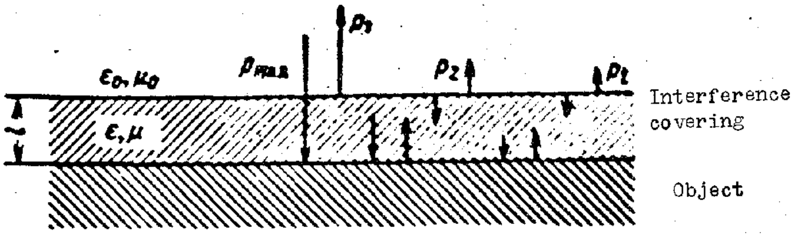


Fig. 9.11. Action of an interference covering.

Equality (9.14) will occur (total reflected field in the direction of the source of the incident wave is equal to zero), if the following conditions are fulfilled:

$$\beta = \ln \frac{1}{|R|}, \tag{9.15}$$

$$l = (2i+1)\frac{\lambda_{\varepsilon,\mu}}{4}. \tag{9.16}$$

Here β is the attenuation factor of a wave for one passage of an absorbing covering in forward and back directions; $|R|$ is the modulus of reflectivity of the covering; ℓ is the thickness of the covering; $\lambda_{\varepsilon,\mu}$ is the wavelength in the substance of a covering with the parameters ε and μ; $i = 1, 2, 3, \ldots$

Conditions (9.15) and (9.16) determine the parameters of the interference covering. It is interesting to note that the interference covering should also possess absorbing properties. Therefore in its composition there are included ferromagnetic materials with admixtures of carbon black as an absorber.

Interference coverings are not so bulky as compared to absorbing coverings. However they have according to their principle of action, less range, which hampers their practical application. More promising are multilayer coverings created on the basis of calculation the properties of absorbing dielectrics and the interference of radio waves reflected from thin metallic films used as a boundary between the dielectric layers.

Interference coverings developed abroad are made basically from a mixture of rubber and carbonyl iron. The English covering of the MX1 type applied in the 3-3.4 cm range has a thickness of 2 mm and a weight of 7 kg/m^2 [63]. The dependence of reflectivity on the length of a normally incident wave is shown in Fig. 9.12.

A characteristic peculiarity of interference coverings is the rather essential dependence of reflectivity on the angle of incidence of the wave. In Fig. 9.13 such a dependence is given for the indicated covering of the MX1 type.

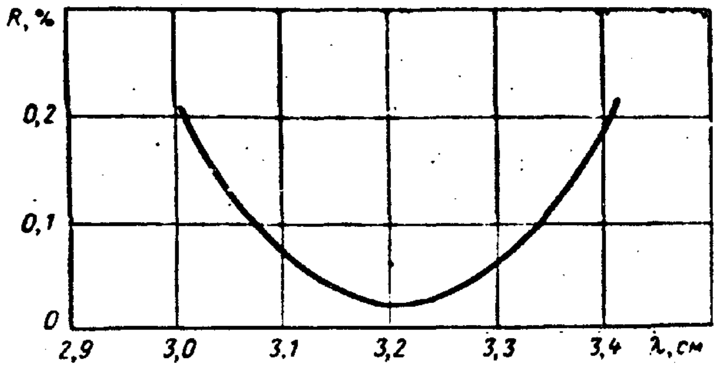


Fig. 9.12. The dependence of reflectivity on wavelength for an interference covering of the MX1 type.

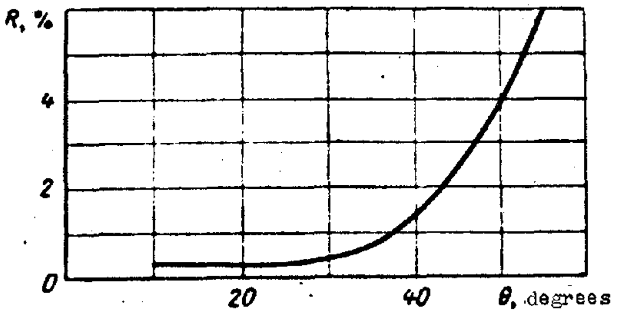


Fig. 9.13. The dependence of reflectivity on the angle of incidence of a wave.

The overall deficience of antiradar coverings of all types is their relatively low range and the large weight of a square meter of covering. The last deficiency is the basic cause of the limited application of coverings in aviation.

In connection with improving the aerodynamic shape of contemporary aircrafts it has become possible to apply covering only on those parts of an object, which give maximum reflection the so-called "brilliant points."

The "brilliant points" on aircraft are mainly joints and sharp functions acting as corner reflectors, air inlets and other apertures, which are considerable (in area) sections of the surface of small curvature with normal incidence on them of an irradiating field, and sharp edges.

A considerable reduction in the power reflected from convex surfaces can be attained by covering only the so-called first Fresnel zone with an absorbing layer.

[Translator's Note: Pages 350-351 of the original document were in part illegible and I have attempted to translate to the best of my ability.]

For convex scattering surfaces with the incidence on them of a plane wave the first Fresnel zone was defined as the section of the scattering surface, included between two parallel planes located at a distance of d = λ/4 and perpendicular to the direction to the source of the radiation; one of the planes is tangent to the dispersing surface (Fig. 9.14). The remaining Fresnel zones (2nd, 3rd, etc.) will be sections of the dispersing surface included respectively between the second and third, the third and fourth, etc. planes parallel to the two earlier mentioned planes and at a distance from each of λ/4.

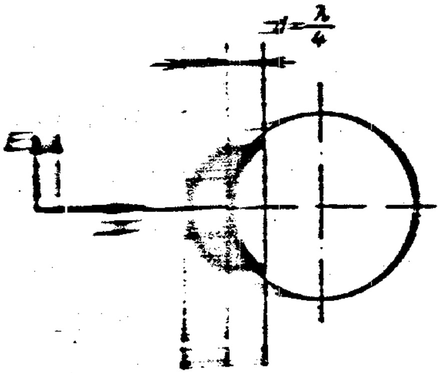


Fig. 9.14. Fresnel zones.

GRAPHIC NOT REPRODUCIBLE

It is natural that when the incoming direction of a radio wave is changed the position of the Fresnel zones is shifted in a corresponding manner. This limits the .... region of application of the method of protection based on the application of coverings within the limits of the first Fresnel zone.

The basis of the main difficulties connected with the application of jamming coverings in aviation is the influence of temperature on their electrical properties and .... characteristics. These difficulties at the present time are being overcome by the application to the surfaces of heat-resistant films.

For surface objects the demands to lower their ESA are less rigid, since the purpose of lowering the ESA in these cases is the camouflaging of important objects (bridges....., etc.) against the background of the surrounding locality. Simultaneously at a certain safe distance from the real targets the false radar reference points (targets) coincide.

For application under surface conditions jamming covers are made in the form of hair, rubber or wooden mats impregnated with a mixture of neoprene ..... rubber) and carbon black. Similar mats with a thickness of several centimeters are able to decrease the power of a reflected signal by 20-50 times. From readily available materials it is possible to use with success coverings made from wet hay and grass.

Antiradar coverings are used under laboratory conditions to eliminate reflections of radio waves from surrounding objects (walls, ceilings, various instruments).

It is necessary to note that even hundred-percent covering of aircraft with antiradar covering cannot ensure their complete camouflage. The fact is that operating engines of aircraft and rockets will form a trail of ionized particles of hot gas. The ionized trail reflecting radio waves is observed on radar screens. Recently it was established that in the flight of supersonic aircrafts there will be formed a readily detectable (by radar) trail even in the case when the engines of this aircraft are not operating.

Control of the Scattering of Radio Waves

An essential reduction of the ESA of a target in principle can be attained by controlling the parameters of the secondary (dispersed) field. The problem of such control is changing the properties of the target, as a reradiating source, in such a degree, so that the minimum reradiated energy is obtained in the necessary direction [66, 71, 82, 83].

One of the methods of control is connecting the overall load to the reflecting object. This method has a certain similarity with above described methods of decreasing ESA with antiradar coverings. However its fundamental difference consists in the fact that in changing the reflecting properties of the target in the examined case there is employed the connecting of the overall load to the local region, the dimensions of which are considerably less than the dimensions of the whole reflecting object. The loaded region in this particular case can be an aperture with concentrated or distributed loads.

In Fig. 9.15 there is depicted target Ц with an aperture of communication s, loaded on the overall load. The target is irradiated by transmitter A, and the reception of the reradiated radio waves is carried out at point B.

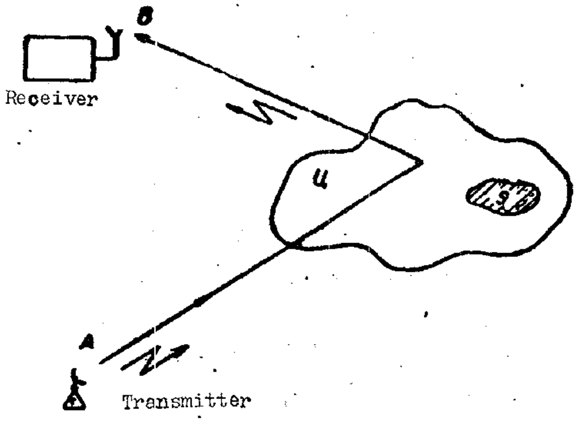


Fig. 9.15. Connecting an overall load to a reflecting object for the purpose of controlling the dispersion of radio waves.

A secondary field at the point of reception B can be represented as the result of the superposition of two fields. One of these is the field of the unloaded body Ц, and the second is the field of the loaded aperture s. It is necessary to note that in view of the small area of aperture s the overall configuration of the target and its area can be considered constant.

The field of aperture scattering s is determined by the form of the aperture and the parameters of load impedance. By the regulating of these parameters it is possible to vary the distribution of amplitude and the phase of the reradiated aperture of the field and, as result of this, to obtain the necessary reduction of the resultant field at the point of reception B.

The relative change of ESA of the loaded target can be estimated by formula [82]

$$\frac{\sigma_{\text{цн}}}{\sigma_{\text{ц0}}}=(1-\alpha^*)^2\left|\frac{Z+\frac{Z_A}{1-\alpha^*}}{Z+Z_A}\right|, \qquad (9.17)$$

where $\sigma_{\text{ц0}}$ is the ESA of the unloaded target; $\sigma_{\text{цн}}$ is the ESA of the loaded target; Z_A is the equivalent overall load from the side of its points of connection in the absence of excitation created by transmitter A; Z is the overall load; α^* is a function of the mutual place of location of the transmitter, receiver and of the form of the target and also of the place of the aperture on the object and the character of the load.

In Fig. 9.16 there is given the qualitative dependence of $\sigma_{\text{цн}}/\sigma_{\text{ц0}}$ on reactive load Z = jx. The minimum ESA of a loaded target corresponds approximately to the reactive load

$$Z=-\mathrm{Im}[Z_A(1-\alpha^*)].$$

The finding of the functions of the form (9.17) for targets with complicated configurations is connected with insurmountable mathematical difficulties. However for bodies with simple form (dipoles, sphere) solutions are obtained in the form of graphs.

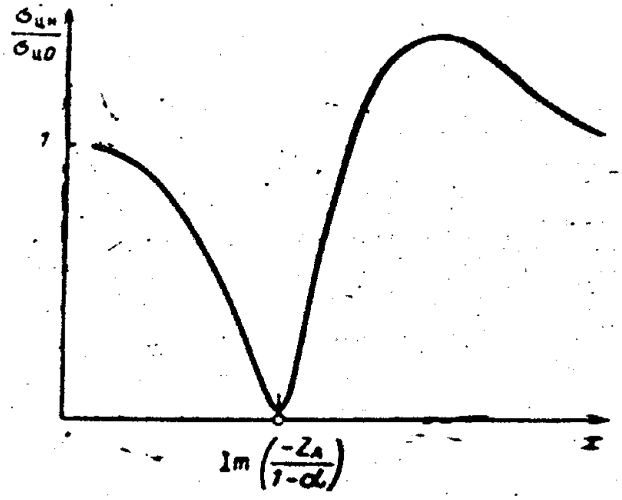


Fig. 9.16. The dependence of the relative value of the ESA of a loaded dipole on the overall load.

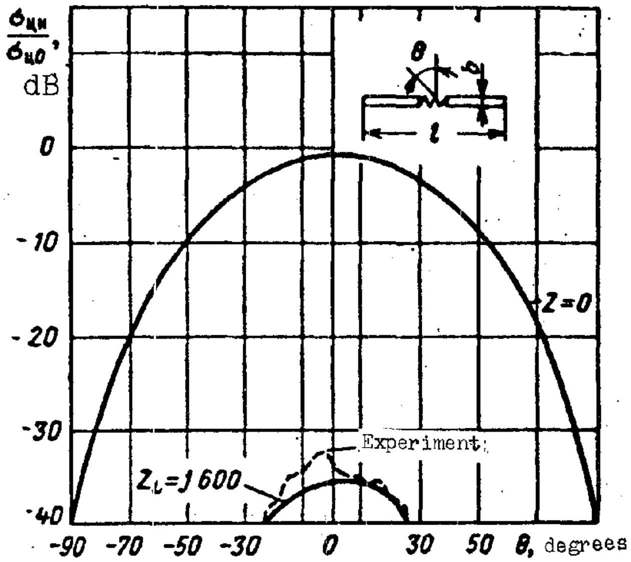


Fig. 9.17. The dependence of the relative value of ESA of a dipole loaded with inductance on the angle of reradiation.

The calculations show that the ESA of a thin dipole by connection to the overall load can be lowered by 20-35 dB. Physically this is explained by the detuning of the dipole introduced by the reactive load.

In Fig. 9.17 there is given the dependence of the relative value of ESA of a loaded dipole on the angle of reradiation θ.

The dipole was loaded with the inductive load $Z_L = j600$. Here there is given the analogous dependence in the case of an unloaded dipole Z = 0. (Length of the dipole $\ell = 0.43\lambda$, thickness $b = 0.0346\lambda$).

A change of the parameters of the overall load can be attained by connecting the concentrated or distributed reactance realized in the form of various cavities (for example, annuluses).

In Fig. 9.18 there are given the relative values of the ESA of a loaded sphere taken experimentally [83]. The overall load is an aperture (annulus); the value and character of the load were regulated by varying the depth of the aperture (annulus). In this case the depth of the aperture (annulus) was changed by replacing the short-circuiting disks.

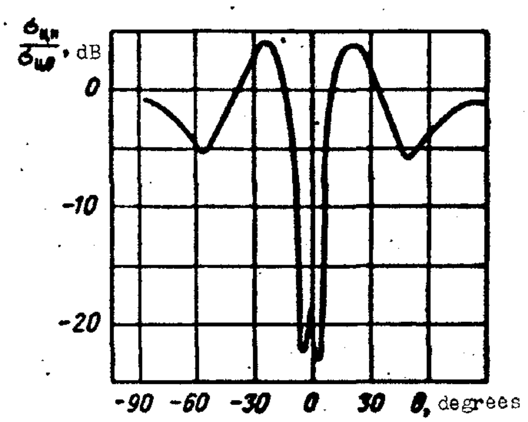


Fig. 9.18. The experimental dependence of the relative value of the ESA of a sphere, loaded on the overall load.

In Fig. 9.19 the geometry of a reflecting sphere is presented.

Experimentally was investigated disk with following parameters;

$$R=42{,}25\ мм,\ d=1{,}6\ мм,\ \theta^\circ=90^\circ,\ \frac{2\pi}{\lambda}r=4{,}28;$$

$$f=5\,136\ \text{GHz}.$$

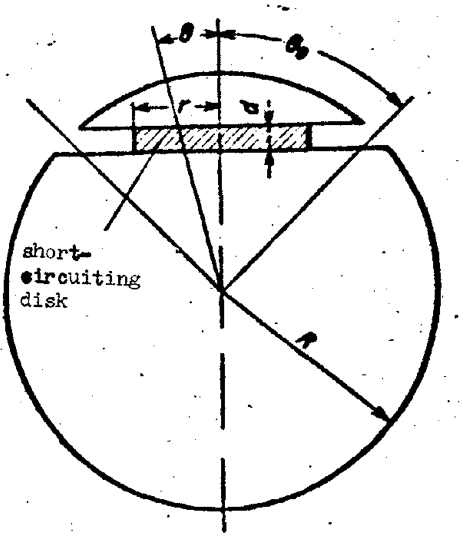


Fig. 9.19. The geometry of a reflecting sphere.

The dependence of the ESA of loaded reradiators on the angle of reradiation (Figs. 9.17 and 9.18) attests to the possibility of reducing the ESA to 20-35 dB.

In practice control of the characteristic of reradiation of an aircraft can be attained with an oscillatory contour created with metallic strips glued or sprayed-on a covering skin [66]. The surface of an aircraft is first covered with insulation material, and then metallic strips are superimposed on it. These strips are oriented and are connected in a different manner (Fig. 9.20), in order to obtain an air capacitor, the capacitance of which slightly depends on the polarization of an incident wave.

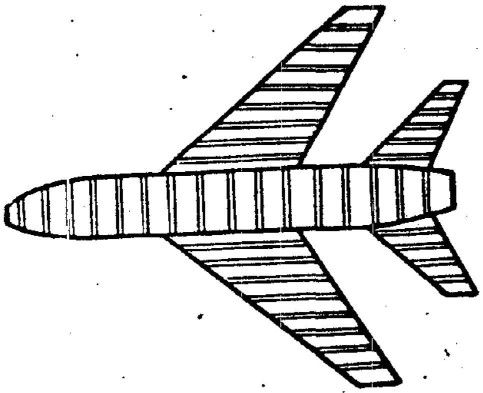


Fig. 9.20. An illustration of the principle of controlling the characteristic of aircraft reradiation.

Air capacitor C_1 is the reactance of the oscillatory contour consisting of inductance L, variable capacitors C_2 and C_3 and resistor R, which is an absorber of electromagnetic energy (Fig. 9.21). Indicator ***И***, connected to the oscillatory contour serves to determine the moment of irradiation of the aircraft. With its help by changing the capacitance of variable capacitors C_2 and C_3 the oscillatory circuit is tuned in resonance with the frequency of the irradiating radar. Resistor R regulates the circuit damping, and thereby the reflectivity of the radio waves from the aircraft.

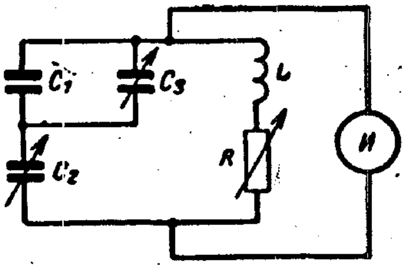


Fig. 9.21. An equivalent diagram of a device controlling the reradiation characteristic of an object.

Considerable interest is being manifested in research to create self-adjusting devices controlling the characteristic of reradiation for the purpose of creating of a corresponding ESA in a definite direction.

CHAPTER 10

ELECTRONIC RECONNAISANCE

10.1 Assignment and Missions of Electronic Reconnaissance

Electronic reconnaissance is a component part of military intelligence. In contrast to all other forms of military intelligence information about the enemy in the case of electronic reconnaissance is obtained by means of an analysis of signals from his electronic devices.

The assignment of electronic reconnaissance is:

– exposure of the system of electronic security of the enemy;
– determination of parameters of electronic means.

Besides electronic reconnaissance, there are also other forms of intelligence with the application of electronic means, for example:

– radar reconnaissance, carried out with help of radar aircraft, for the purpose of exposing enemy objects;
– television reconnaissance, carried out with help of aircraft and other television devices.

Electronic reconnaissance is one of the basic methods for obtaining information about parameters and disposition of hostile electronic devices and their coordinates.

With the help of electronic reconnaissance the following missions are fulfilled:

- carrier frequency is determined;
- direction of arrival of wave is measured (position of electronic device);
- the type of electronic device reconnoitered is identified (detection radar, Gun-laying radar [SON], radio link, etc.);
- measurement (appraisal) is made of parameters of reconnoitered electronic devices (frequency of repetition, duration of pulses, structure of lateral antenna lobes, polarization, form of modulation, etc.);
- a recording is made of reconnaissance findings in a memory unit for subsequent analysis.

Results of electronic reconnaissance are used for making a decision concerning the selection of methods of jamming in a developing combat situation, namely:

- to establish the necessity for suppression of exposed electronic devices;
- to determine the detailing of forces and means for jamming;
- to select the optimum operating conditions for jamming transmitters (form of interferences, form of interference modulation, moment of switching on and switching off of jamming transmitters).

10.2. Application of the Queueing Theory of the Solving of Problems of Electronic Reconnaissance

The high rate of saturation of a contemporary Antiaircraft Defense [PVO] system with electronic devices, especially pulse radar, leads to the necessity to examine the problem of electronic reconnaissance within the bounds of the queueing theory. The queueing theory has been developed quite completely, however, in electronic reconnaissance it has just started to be applied only very recently. Elements of the queueing theory in a form which is convenient for application in electronic reconnaissance are expounded in books by

Ye. S. Venttsel' [15, 84]. As was already noted earlier, the mission of electronic reconnaissance includes detection and determination of parameters of the corresponding electronic devices by means of reception and analysis of their signals. Reception of signals and their analysis can be examined as a unique service.

Application of the queueing theory makes it possible to arrive at the solution to the following most important problems of electronic reconnaissance.

1. Based on assigned parameters of flow of signals arriving at the input of the reconnaissance device and assigned probability of reconnaisance, to determine the minimum necessary number of channels of reconnaissance and the most permissible value of average time of treatment of a signal received in one channel.

2. For the assigned reconnaissance device and probability of electronic reconnaissance, to determine the maximum permissible number of electronic devices which can be reconnoitered with its help. This in turn makes it possible to select an effective altitude of flight for the reconnaissance plane in order to bring into conformity the carrying capacity of the reconnaissance station with the flow of transmitted signals.

Application of the queueing theory is possible in principle if the characteristics of flow of demands for service are known, in this case the flow of radio signals from the reconnoitered devices, and the characteristics of the actual means of servicing — the reconnaissance equipment.

One of the basic characteristics of a radio reconnaissance unit as a service device is the time of service (time for reception and analysis of signal). In view of the presence of inherent noises in a radio reconnaissance device and random external influences on the receiver, the time of service will be, generally speaking, a random variable. Actually the probability of electronic reconnaissance, determined by the probability of the correct detection of the signal from the electronic device being reconnoited in the noises and the probability of identification of form (P_p), as a function of time of reconnaissance (time of service t_p) is represented by the curves shown in Fig. 10.1. The parameter of the set of curves can be, for example, average time of service or the ratio of power of reconnoitered signal to power of noise. Hence the correctness of the opinions expressed on the random nature of time of reconnaissance becomes evident. The cited curves also makes it possible to consider the temporary nature of operation of the reconnoitered electronic devices, thus, reducing the assumed time of service.

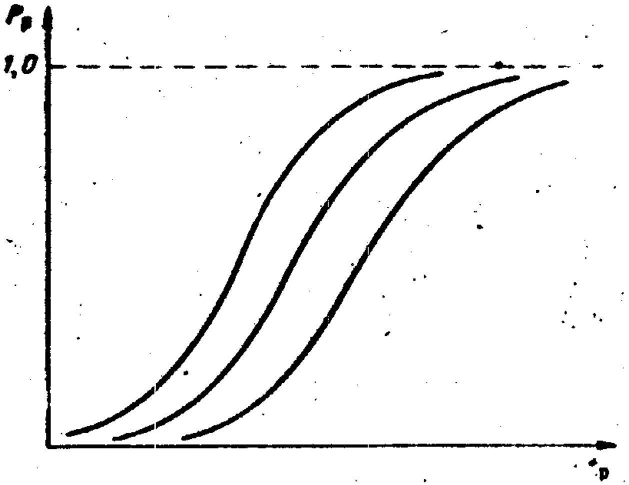


Fig. 10.1. Dependence of probability of electronic reconnaissance on the time of reconnaissance.

The next characteriste of the queueing system is the time of waiting in line. This time in general is also a random variable. In electronic reconnaissance the waiting time is determined by the time of operation of the reconnoitered electronic device, i.e., the time of stay of the reconnoitered device in the system of service (reconnaissance). Here the "serviced" signals can at any moment of time not only retire from the line, but also interrupt the process of service, without waiting for its termination.

Two basic classes of queueing systems are distinguished — a system of service with rejections and a system with delay. In systems of service with rejections a request arriving at a time when the system is occupied is not serviced either at that moment of time or subsequently. In reference to electronic reconnaissance this means that that particular device will not be detected if the signal generated by it enters the queueing (reconnaissance) system at that moment when it is processing a signal from another electronic device. In practice the noted circumstance can take place in the case of reconnaisance of a momentarily operating electronic device (for example, radio links for transmission of single commands and others), and especially in the case of reconnaissance which is conducted for creation of spot jammings. A considerable number of electronic means, especially detection and guidance radar and also target designation radar, operate for a prolonged time, therefore the problem of reconnaissance of such devices should be examined in the plan of the queueing theory with delay.

Before switching to the derivation of fundamental equations for queueing systems, we will dwell on the characteristics of standard flows of signals which are subject to reconnaissance, and a general consideration of data processing in reconnaissance devices. Parameters of the flow of signals depend due to a considerable degree on the theater of military operations, altitude of flight of the aircraft which is carrying out the reconnaissance, and the sensitivity of the reconnaissance device. For the assigned theater of military operations, altitude of flight, and reconnaissance device the flow of signals on the input of the reconnaissance system can be considered with sufficient accuracy as the simplest (stationary Poisson).

We will show this in an example of electronic reconnaissance of a system of pulse detection and guidance radar. In case, when synchronization of radar systems is absent, the regular sequences of pulses of separate radars will superimpose on one another in a random manner. In practice it is sufficient to have an overlapping in time of 4-5 nonsynchronized pulse sequences in order to obtain the simplest flow.

As is known [84], a flow is called the simplest if it is stationary and ordinary, and in it there is no residual effect.

Flow is called stationary if the average flow density of signals does not depend on time, i.e., in other words, the probability of a certain number of signals hitting in an interval of time τ depends only on the length of this interval and does not depend on its position on the axis of time.

Flow is called ordinary if the probability of simultaneous hit of two and more signals in a small interval of time is a value of a second and higher order of smallness as compared to the probability of a hit of one signal. On the basis of the formulated condition the probability of arrival of one signal during the time Δt is equal approximately to $\lambda\Delta t$, where λ – flow density of pulses. The probability of arrival of two signals is accordingly proportional to Δt^2, etc.

A residual effect is absent in flow if the number of signals, arriving in a given interval of time, does not depend on what number of them arrived in other, nonoverlapping, intervals of time. In other words, the probability of arrival of a signal at a given moment of time does not depend on whether or not the arrival of some signal took place in all the preceding moments of time.

Flow density λ, in general, in process of reconnaissance is changed in connection with the fact that there is a change in the number of stations sweeping the reconnaissance device. Therefore, strictly speaking, under conditions of reconnaisance from an aircraft the input flow of signals will not be the simplest, i.e., in practice a condition of stationarity is not fulfilled, while the other two conditions in first approximation are fulfilled; this makes it possible to view the real flow of signals as transient Poisson flow. Considering the slow nature of change of $\lambda(t)$ in time, for tentative calculations it is possible to consider that $\lambda(t) = \text{const}$.

In many electronic reconnaissance stations a multistage treatment of sequence of signals takes place, as a result of which after every stage of treatment a portion of the signals is screened. For example, this can take place at the expense of filtration on high frequency.

To a considerable degree the recording device receives a rarefied flow of signals, which with great reason can be related to Erlang flow of the corresponding order. Nevertheless, considering the approximate nature of the consideration, we will not investigate the process of passage of a sequences of signals in a reconnaissance device by stages. For our purpose it is sufficient to examine the electronic reconnaissance station as a single installation, characterized by time of service and carrying capacity, on the input of which the simplest flow of signals arrives.

As already was noted above, one of the basic characteristics of a reconnaissance device as a means of service is the time of service. In order to show more clearly how this time is determined, and also to comprehend the basic principles of identification of the form of electronic means in the electronic reconnaissance device, we will examine in general form the system of data processing in a reconnaissance device. The basic problems of electronic reconnaissance are reduced to the detection of signals from an electronic device, identification of the type of device revealed, and an appraisal of its basic parameters. By detection of signals in this case is understood the interception by a reconnaissance device of signals from a particular electronic facility. The process of interception of signals, in general, requires a certain time; as a rule, the longest time is required for identification of form.

The process of electronic reconnaissance can be presented in the form of two basic operations. The first operation ensures conversion of the multitude of input signals into a multitude of parameters and criteria, characterizing the forms of facilities being reconnoitered. The second operation includes correspondence of groups (subsets) of parameters (criteria) with concrete forms of electronic facilities. This operation can also be appraisal of parameters of the reconnoitered facilities.

Every electronic facility is determined clearly by a certain totality of independent parameters (carrier frequency, angle of arrival of radio waves at a given point of observation, width of beam, polarization, power, pulse duration, frequency of following of pulses, angular velocity of rotation of antenna, and others). With help of the reconnaissance device these parameters will be converted either to a form which is convenient for observation by an operator, who records them in an appropriate manner, or to form which is convenient for recording on photographic film or magnetic tape, or into voltages and currents, which in turn are coded by a binary code and are recorded in the form of numbers of a binary system in memory of an electronic digital computer.

Each electronic means of a given class, characterized by an assigned number out of n independent parameters, can be linked to a vector of criteria in n-dimensional metric space. As a base of space it is expedient to select the stated n independent parameters which are recorded by the reconnaissance device. In order to construct a vector of criteria corresponding to the form of the reconnoitered electronic means a comparatively great deal of time is required in the process of reconnaissance. The greater the time expended on the formation of vector of criteria, then, generally speaking, the greater the probability of a correct decision about the form of the electronic means. At the same time an increase of time for identification of form decreases the carrying capacity of the electronic reconnaissance station and thereby increases the probability of losing the object being reconnoitered.

Besides parameters, electronic device has certain specific features which are peculiar to it. The most characteristic features of radar are form of pulse, form of antenna radiation pattern, and the fine structure of spectrum of sequence of signals. It is possible to affirm, for example that the fine structure of antenna radiation patterns of antennas from two radars of the same class and same range, and found approximately identical conditions, will be different. Each of the antenna radiation patterns will have its own characteristic "blips," which after the corresponding analysis makes it possible to definitely state to which radar the given antenna radiation pattern belongs. An analogous picture also takes place for the fine structure of form and spectrum of pulse signals. In other words, signals from each electronic device have their own characteristic imprint.

This makes it possible in a number of cases to be limited, for example, to an analysis of form of only one pulse for making the final solution, which essentially reduces the time of electronic reconnaissance.

We will limit ourselves to a consideration of two of the most characteristic systems for electronic reconnaissance:

- single-channel queueing system with rejections;

- multichannel system with limited waiting time.

Initially we will derive the equations for the simplest case of a single-channel system.

Single-Channel System of Electronic Reconnaissance with Rejections

A single-channel system practically exists when use is made of radio reconnaissance receivers with very rapid or very slow retunings. Single-channel in the sense of the queueing theory will be a reconnaissance device constituting a total of several dozen straight (amplification) receivers, each of which ensures reception of pulse signals in a comparatively narrow frequency range (system of simultaneous investigation on carrier frequency).

Carrier frequency of the reconnoitered radar is determined approximately with an accuracy of half a transmission band of a high-frequency filter of the corresponding receiver from the stated aggregate. Average time of service by one receiver is equal to several periods of following of pulses from the radar in question.

The arrangement for electronic reconnaissance, similar to the arrangement for queueing, cannot be called multichannel in its full meaning, inasmuch as separate channels are not interchangeable. Each of the receivers of the aggregate services only a radar of a given sub-band, and it cannot service radar signals of another band. In the theory of queueing a multichannel system is considered a system in which each channel which is free at a given moment of time can service any signal belonging to the flow of signals being serviced. Actually in this case we should examine as many independent flows of signals as there are independent receivers. In order that an electronic reconnaissance receiver can be considered a multichannel device in the sense of the queueing theory, it is necessary that each of its channels be able to service any signals of an assigned class.

We will consider that a Poisson flow of pulses arrives at the input of a single-channel reconnaissance receiver. Initially, for community we will not require a stationarity of flow of signals, however we will demand that it be ordinary and without a residual effect.

We will determine the probability $P_0(t)$ that the queueing system will be free at the time t of arrival of the reconnoitered signal. In order to derive the corresponding equations for $P_0(t)$, it is necessary to know the probability of arrival of the signal during an assigned time with a Poisson flow of signals and the law of distribution of time for servicing in the system.

Probability $P_n'(t)$ of arrival of n pulses during the time t in a Poisson flow will be determined on the basis of the following reasonings. Due to the absence of residual effect in the flow of signals and its ordinariness, the arrival of n pulses during the time $t + \Delta t$ can take place as a result of the approach of the following two incompatible events:

n signals will arrive during the time t and none will arrive during the time Δt;

(n - 1) signal will arrive during the time t and one signal will come during the time Δt.

If the flow density of signals is designated by $\lambda(t) = \lambda$ (λ – number of signals arriving in a unit of time), then the probability of arrival of one signal during the time Δt will equal $\lambda\Delta t$. Accordingly the probability that the signal will not arrive in the interval of time Δt is equal to $(1 - \lambda\Delta t)$.

Thus, the following recurrence relation is obtained for $P_n'(t + \Delta t)$:

$$P'_n(t+\Delta t)=P'_n(t)(1-\lambda\Delta t)+P'_{n-1}(t)\lambda\Delta t. \tag{10.1}$$

This formula is correct for all $n \neq 0$. If $n = 0$, then the event which interests us can approach by a unique method, namely both during time t and during time Δt not one signal will arrive, i.e.,

$$P'_0(t+\Delta t)=P'_0(t)(1-\lambda\Delta t). \tag{10.2}$$

It follows from this that

$$\frac{P'_0(t+\Delta t)-P'_0(t)}{\Delta t}=-\lambda P'_0(t).$$

Passing on to the limit at $\Delta t \to 0$, we obtain the following differential equation for $P_0'(t)$ – the probability that during the time t from a Poisson flow of signals with density $\lambda = \lambda(t)$ not one signal will arrive:

$$\frac{dP'_0(t)}{dt}=-\lambda P'_0(t). \tag{10.3}$$

Differential equation (10.3) is solved under the following evident initial condition $P_0'(0) = 1$. (In the initial moment of time $t = 0$ the probability absence of a signal on the input is equal to a unit).

The solution of equation (10.3) under the assigned initial condition, as is known, can be presented in the following way:

$$P'_0(t)=e^{-\lambda t}. \tag{10.4}$$

With the help of recurrence relation (10.1) we determine the probability, which we are interested in, of $P'_n(t)$ for Poisson flow of signals

$$P'_n(t)=\frac{(\lambda t)^n}{n!}e^{-\lambda t}. \qquad (10.5)$$

The most convenient approximation of the law of distribution of queueing time is the exponential law. Strict consideration shows that selection of another approximation of the law of distribution does not significantly affect the quantitative results, however, to a considerable degree it complicates the equations and their investigation. On the force of the accepted approximation of the law of distribution of queueing time the probability that the time for handling the signal will be less than t is equal to

$$P(\tau<t)=1-e^{-\mu t}. \qquad (10.6)$$

Here $\mu=\frac{1}{\bar{t}_{об}}$; $\bar{t}_{об}$ — average time for handling.

In principle a single-channel queueing system can be found in two states:

B_0 – system is free and can handle a signal arriving at the given moment of time. Probability of this state $P(B_0) = P_0(t)$;

B_1 – system is occupied in handling a signal which it accepted earlier. Any signal arriving at that instant turns out to be generally unserviced, i.e., there is an omission of the reconnoitered facility.

Probability of such a state $P(B_1) = P_1(t)$. Inasmuch as the system of events B_0 and B_1 is complete, then

$$P_0(t)+P_1(t)=1. \qquad (10.7)$$

We will determine the probability of the system remaining in state B_0 at the moment of time $t + \Delta t$, if at the instant t it was found in any of the states which are possible for it under the given conditions. The event which interests can set in by the following two incompatible methods.

1. At the moment of time t the system was in state B_0 and during the time Δt not one signal arrived for handling. Probability of the joint occurrence of these events is equal to

$$P_0(t)e^{-\lambda\Delta t} \approx P_0(t)(1-\lambda\Delta t).$$

2. At instant t the system was occupied, and in the interval of time Δt it was cleared, i.e., after this small interval of time servicing is completed. On the force of invariance of the exponential law of distribution, expressed in the fact that the law of distribution of remaining time for handling will also be exponential independent of how much time before this the given service lasted, the probability that handling will be completed during time Δt equals $(1-e^{-\mu\Delta t})$. Probability of joint occurrence of two events is equal to

$$P_1(t)(1-e^{-\mu\Delta t}) \approx P_1(t)\mu\Delta t.$$

Full probability of occurrence of the event which interests is equal to

$$P_0(t+\Delta t) = P_0(t)(1-\lambda\Delta t) + P_1(t)\mu\Delta t$$

or

$$\frac{P_0(t+\Delta t)-P_0(t)}{\Delta t} = -\lambda P_0(t) + \mu P_1(t).$$

Passing on to the limit at $\Delta t \to 0$, we obtain the first differential equation for a single-channel queueing system

$$\frac{dP_0}{dt} = -\lambda P_0 + \mu P_1. \qquad (10.8)$$

The second differential equation, connecting P_0 and P_1, can be obtained by determining the probability of stay of the system at instant $t + \Delta t$ in state B_1.

By the moment of time $t + \Delta t$ the system can arrive at state B_1 by two routes.

First route – at instant t the system of reconnaissance handled a signal and in the time Δt service was completed. This can take place with a probability of

$$P_1(t)e^{-\mu\Delta t} \approx P_1(t)(1-\mu\Delta t).$$

Second route – at instant t the system of reconnaissance was free, but during the interval of time Δt signal for service arrived. Such combination of events can occur with the probability

$$P_0(t)(1-e^{-\lambda\Delta t}) \approx P_0(t)\lambda\Delta t.$$

Full probability of employment of the system at the instant $t + \Delta t$ is equal to

$$P_1(t+\Delta t) = P_1(1-\mu\Delta t) + P_0\lambda\Delta t.$$

Hence after evident conversion and passage to the limit we obtain the desired differential equation for $P_1(t) = P_1$:

$$\frac{dP_1}{dt} = -\mu P_1 + \lambda P_0. \qquad (10.9)$$

The probability that at the instant $t = 0$ there will be no signal and, consequently, the system will be free, is equal to a unit. This circumstance determines the following initial conditions of system of equations (10.8) and (10.9):

$$P_0|_{t=0} = 1, \quad P_1|_{t=0} = 0.$$

Using equality (10.7) and equation (10.8), it is possible to write the equation for determination of P_0

$$\frac{dP_0}{dt} = -(\lambda+\mu)P_0 + \mu. \qquad (10.10)$$

The solution of equation (10.10) in case $\lambda = \lambda(t) = \text{const}$ and the above mentioned initial conditions has the form

$$P_0 = \frac{\mu}{\lambda+\mu} + \frac{\lambda}{\lambda+\mu} e^{-(\lambda+\mu)t}. \qquad (10.11)$$

The solution obtained makes it possible to determine the basic parameters of the system of electronic reconnaissance in that case when it can be presented in the form of a single-channel queueing system with rejections (single search, reconnaissance of electronic means which operate for an extremely limited time).

Relative carrying capacity of a system of radio reconnaissance with series single search is equal to P_0. Really, by definition the relative carrying capacity of a system is the ratio of average number of signals handled to the average number of signals arriving at the input of the reconnaissance device. Inasmuch as a request can be handled only in that case when the system is free, i.e., with probability P_0, then numerical values of relative carrying capacity and P_0 coincide.

The value P_0 also determines the probability of detection of the assigned electronic device, inasmuch as a necessary condition for interception of signals from the assigned device is the readiness of the system to handle them. Necessary and sufficient conditions for interception at a given moment of signals from the designated device are readiness of the system to handle them and presence at that instant of the required signal. Probability of the latter event $P^{(n)}$ for systems of continuous sweep with directional antennas is approximately equal to

$$p^{(n)} = \frac{\theta_{0,5}}{2\pi},$$

where $\theta_{0,5}$ — width of beam of antenna.

For pulse systems

$$p^{(n)} = \frac{\theta_{0,5}}{2\pi Q},$$

where Q – porosity.

In order to obtain the probability of electronic reconnaissance $P^{(1)}_{\text{ртi}}$ it is still necessary to consider the probability of correct identification of form of the asigned electronic facility $P_{\text{рэс}}$. Hence the probability that at a given moment complete electronic reconnaissance of the assigned device will be carried out is equal to

$$P^{(1)}_{\text{iтр}} = P_0 P^{(n)} P_{\text{рэс}}.$$

The probability of electronic reconnaissance of an assigned device, represented by its own signals, in the common Poisson flow of demands is equal to

$$P^{(1)}_{\text{ртр}} = P_0 P_{\text{рэс}}.$$

We recall once again that here the talk concerns determination of the probability of electronic reconnaissance by single-channel reconnaissance devices with rejections.

In the established queueing regimen ($t \to \infty$)

$$P_0 = \frac{\mu}{\lambda + \mu}. \tag{10.12}$$

This formula is convenient to use for determination of necessary average time of handling, ensuring a rated value of probability of reconnaissance P_0 at a given Poisson flow of signals with a density λ = const:

$$\bar{t}_{\text{об}} = \frac{1 - P_0}{\lambda P_0}. \tag{10.13}$$

Absolute carrying capacity of a single-channel system of reconnaissance (number of signals handled on the average per unit of time) in the established regimen is equal to

$$q = P_0 \lambda$$

or

$$q = \frac{\lambda\mu}{\lambda + \mu}. \tag{10.14}$$

We will assume that each of n radars of a certain sub-band of waves, serviced by a particular electronic reconnaissance station, sweeps an aircraft with electronic reconnaissance equipment over intervals of time which are on the average identical for all n radars and are equal to the mathematical expectation for the period of rotation of the radar antennas. Mathematical expectation of the period of rotation of the antenna of one radar is determined by the ratio

$$T_{\lambda}=\frac{60}{N},$$

where N — average number of turns of the antenna per minute.

Accordingly the average time between signals entering the electronic reconnaissance station equals

$$T_{c}=\frac{60}{Nn}.$$

Hence the sought value of flow density λ of reconnoitered signals will be determined by the formula

$$\lambda=\frac{Nn}{60}.$$

We will accept the average time of handling as equal to the mathematical expectation of sweep time of the aircraft by the major lobe of the antenna radiation pattern in a horizontal plane

$$t_{cp}=\frac{\theta_{0.5}}{6N},$$

where $\theta_{0.5}$ — width of beam of radar antenna on half power.

Considering the beam width for all N radars identical, we find

$$\mu=\frac{1}{t_{cp}}=\frac{6N}{\theta_{0.5}}.$$

We are interested in the probability that the incoming signal will be handled. This takes place, if the electronic reconnaissance station is free. Probability of the latter event is

$$P_{0}=\frac{\mu}{\mu+\lambda}.$$

Carrying out substitution of the previously obtained values of λ and μ, we obtain the final formula, determining the probability of electronic reconnaissance in the assigned conditions

$$P_0 = \frac{360}{360 + n\theta_{0,5}}.$$

Let us assume that n = 30 radars, $\theta_{0.5} = 1°$, then, accordingly, $P_0 \approx 0.92$.

Multichannel System of Electronic Reconnaissance in the Case of a Limited Waiting Time for Signals at the Input of the Receiver

Without dwelling on the multichannel radio reconnaissance receiver, which is equivalent in principle of functioning to a multichannel system of queueing with rejections, let us turn to a study of multichannel devices used to carry out electronic reconnaissance of electronic devices which are working continuously during certain finite time intervals, in general distributed by random law. The conditions of electronic reconnaissance examined below are typical.

For convenience of calculations we will consider the law of distribution for time of continuous operation of the reconnoitered electronic device and the law of distribution for time of handling as exponential. This means that the probability of continuous operation of the reconnoitered electron device during time t is equal to

$$P(\tau < t) = 1 - e^{-\chi t}.$$

Here

$$\chi = \frac{1}{\bar{t}_{ож}};$$

$\bar{t}_{ож}$ — average time of continuous operation of reconnoitered electronic device, average waiting time.

Accordingly the density of distribution of time of continuous operation is equal to

$$\frac{dP(\tau < t)}{dt} = \chi e^{-\chi t}.$$

In the theory of queueing an analog of such a type of arrangement for reconnaissance is a system with a limited waiting time for the facility being serviced in the queueing system. Here the discussion concerns waiting namely in the queueing system, and not in the line serviced by an individual channel. Let us assume that the system of reconnaissance has n independent channels, each of which can handle any signal from the assigned Poisson flow of signals at the input. The following states of the electronic reconnaissance system are possible.

A_0 – signals are absent at the input of the system, all the channels are free;

A_1 – at the input of the system one signal has arrived and is handled in one of the channels, the other n - 1 channels are free. There are no lines;

A_i – at the input of the electronic reconnaissance system i signals arrived and all of them are handled by i channels, arbitrarily selected from n. There are no lines;

A_n – n signals are handled and there is no line at the input of the reconnaissance system;

A_{n+1} – all n channels are occupied in handling and at the input of the electronic reconnaissance system there is one signal. Average time of stay of a signal in the system

$$\bar{t}_{\text{ож}} = \frac{1}{\chi};$$

A_{n+k} – all channels are occupied in handling and, furthermore, at the input there are k signals, waiting for a limited time for handling.

The number of possible conditions of the system is infinitely great on the force of unlimitedness in the time of flow of the reconnoitered signals. Assuming the Poisson nature of flow of signals at the input of the reconnaissance system, we determine the probability that at the instant t + Δt the system will be found accordingly in each of these conditions, if in the preceding fixed instant t it was found in any of the states which are possible for it under the given conditions. By the time t + Δt condition A_0 may be reached by two incompatible routes:

— at instant t the system was free (condition A_0) and in the time Δt not one signal arrived;

— at instant t the system was found in condition A_1 and during time Δt handling was completed.

These reasonings make it possible to write the following equality:

$$P_0(t+\Delta t)=P_0(t)(1-\lambda\Delta t)+P_1(t)\mu\Delta t.$$

Hence we obtain a differential equation for $P_0(t)$:

$$\frac{dP_0}{dt}=-\lambda P_0(t)+\mu P_1(t).$$

Let us turn to a determination of the probability of condition A_i in a multichannel system of reconnaissance. At the instant $t + \Delta t$ the reconnaissance system can arrive at condition A_i as a result of the occurrence of the following three incompatible events:

— at the instant t the system was found in condition A_i, during the time Δt not in one of i channels was handling completed and during the same time not one new signal arrived; probability of joint occurrence of all these three events is equal to

$$P_i(t)(e^{-\mu\Delta t})^i e^{-\lambda\Delta t}\approx[1-(\lambda+i\mu)\Delta t]P_i(t);$$

— at the instant t the system handled i - 1 signals (condition A_{i-1}), during the time Δt one more signal arrived for handling, and service was not completed for one of the i - 1 signals accepted by the radio reconnaissance device; probability of joint occurrence of these events, with an accuracy up to an infinitesimal higher order than Δt, is equal to

$$P_{i-1}(t)(1-e^{-\lambda\Delta t})e^{-(i-1)\mu\Delta t}\approx P_{i-1}(t)\lambda\Delta t;$$

— in instant t the system was found in condition A_{i+1}, during the time Δt in one of i + 1 channels handling was completed and not one signal arrived at the input of the reconnaissance receiver; accordingly the probability of joint occurrence of these three events

$$P_{i+1}(t)C'_{i+1}(1-e^{-\mu\Delta t})e^{-\lambda\Delta t}\approx P_{i+1}(t)(i+1)\mu\Delta t.$$

Full probability of the occurrence of the event which interests us

$$\begin{aligned}P_i(t+\Delta t)=P_i(t)[1-(\lambda+i\mu)\Delta t]+\\+P_{i-1}(t)\lambda\Delta t+P_{i+1}(t)(i+1)\mu\Delta t.\end{aligned}$$

Hence after evident conversions and transition to the limit we obtain differential equation

$$\frac{dP_i(t)}{dt}=\lambda P_{i-1}(t)-(\lambda+i\mu)P_i(t)+(i+1)\mu P_{i+1}(t).$$

We will determine the probability of an A_n condition for the system at the instant t + Δt. It can take place as a result of the occurrence of the following three incompatiable events:

– at the instant t the system was in condition A_{n-1}, during the time Δt one signal arrived for handling, and not one of n - 1 signals was serviced;

– at the instant t the system was in condition A_n, during the time Δt not one signal arrived, and handling had not been completed of any of the earlier accepted n signals;

– at the instant t the system was found in condition A_{n+1}, i.e., n signals were handled and one had arrived at the input without being accepted for handling (stood in line). During the time Δt no additional signals arrived at the input for handling and the system passes into condition A_n either due to the fact that the signal is accepted for handling by one of the freed channels, or in view of the fact that the signal leaves the system without being serviced.

Probabilities for the two first events were determined above. Probability of the latter (third) event will be determined by the following sum of probabilities of particular events:

$$\begin{aligned}-\quad P_{n+1}(t)e^{-\lambda\Delta t}C'_n(1-e^{-\mu\Delta t})+\\+P_{n+1}(t)e^{-\lambda\Delta t}(1-e^{-\chi\Delta t})\approx P_{n+1}(t)(n\mu+\chi)\Delta t.\end{aligned}$$

The differential equation for P_n taking into account the probabilities of the two first incompatible events from general number determining condition A_n, has the form

$$\frac{dP_n(t)}{dt} = \lambda P_{n-1}(t) - (\lambda + n\mu) P_n(t) + (n\mu + \chi) P_{n+1}(t).$$

We will determine, finally, the probability that at the instant $t + \Delta t$ the system will be in condition A_{n+k}. Condition A_{n+k} can be achieved by the following incompatible routes:

– at the instant t the system was in condition A_{n+k-1}, and during the time Δt exactly one request arrives;

– at the instant t the system was in condition A_{n+k}, during the time Δt not one signal arrived at the input, handling was not completed for any of n signals, and not one of k signals got out of line;

– at the instant t the system was in condition A_{n+k+1}, during the time Δt no new signals arrive, and the system passes into condition A_{n+k} either due to the releasing of one of the channels, or in view of the departure of one of the k + 1 signals. The full probability of occurrence of event A_{n+k} at the instant $t + \Delta t$ will accordingly be equal to

$$\begin{gathered} P_{n+k}(t + \Delta t) = P_{n+k+1}(t)(1 - e^{-\lambda\Delta t}) e^{-n\mu\Delta t} e^{-(k-1)\chi\Delta t} + \\ + P_{n+k}(t) e^{-\lambda\Delta t} e^{-n\mu\Delta t} e^{-k\chi\Delta t} + \\ + P_{n+k+1}(t) e^{-\lambda\Delta t} [n(1 - e^{-\mu\Delta t}) + k(1 - e^{-\chi\Delta t})] \end{gathered}$$

or

$$\begin{gathered} P_{n+k}(t + \Delta t) \approx P_{n+k+1}(t) \lambda\Delta t + \\ + P_{n+k}(t) [1 - (\lambda + n\mu + k\chi)] \Delta t + \\ + P_{n+k+1}(t) [n\mu + (k+1)\chi] \Delta t. \end{gathered}$$

Hence after passage to the limit we obtain the desired differential equation

$$\begin{gathered} \frac{dP_{n+k}(t)}{dt} = \lambda P_{n+k+1}(t) - (\lambda + n\mu + k\chi) P_{n+k}(t) + \\ + [n\mu + (k+1)\chi] P_{n+k+1}(t). \end{gathered}$$

Thus we obtained the following infinite, but countable system of differential equations for probabilities of conditions of a multichannel system which is carrying out electronic reconnaissance of electronic devices which are operating for a limited time:

$$\begin{gathered}\frac{dP_0}{dt}=-\lambda P_0(t)+\mu P_1(t),\\ \frac{dP_i(t)}{dt}=\lambda P_{i-1}(t)-(\lambda+i\mu)P_i(t)+(i+1)\mu P_{i+1}(t),\\ \cdots\cdots\cdots\cdots\\ \frac{dP_n(t)}{dt}=\lambda P_{n-1}(t)-(\lambda+n\mu)P_n(t)+(n\mu+\chi)P_{n+1}(t),\\ \frac{dP_{n+k}(t)}{dt}=\lambda P_{n+k-1}(t)-(\lambda+n\mu+k\chi)P_{n+k}(t)+\\ +[n\mu+(k+1)\chi]P_{n+k+1}(t),\end{gathered} \tag{10.15}$$

Initial conditions of the resulting system of equations (10.15) have the form

$$P_0(t)|_{t=0}=1,\ P_j(t)|_{t=0}=0\ (j=1,\ 2,\dots,n+k,\dots).$$

In the established regimen of electronic reconnaissance all probabilities P_j ($j = 0, 1, \dots, n + k$) can be considered constants; accordingly the system of differential equations will be converted into a system of algebraic equations:

$$\begin{gathered}-\lambda P_0+\mu P_1=0,\\ \lambda P_{i-1}-(\lambda+i\mu)P_i+(i+1)\mu P_{i+1}=0,\\ \lambda P_{n-1}-(\lambda+n\mu)P_n+(n\mu+\chi)P_{n+1}=0,\\ \lambda P_{n+k+1}-(\lambda+n\mu+k\chi)P_{n+k}+\\ +[n\mu+(k+1)\chi]P_{n+k+1}=0.\end{gathered} \tag{10.16}$$

To this system of equations it is necessary yet to add the evident equality

$$\sum_{j=0}^{\infty}P_j=1. \tag{10.17}$$

A signal, arriving at the input of the reconnaissance system, may be detected and treated (serviced) or can be passed through, it can leave the line without being serviced. If one were to designate by P_p the probability that the incoming signal or group of signals will be accepted and treated in receiver of the electronic reconnaissance device, and by $P_{\text{н}}$ — the probability of passing the signal (probability that the accepted signal will leave the line without being serviced), the always

$$P_{\text{р}} + P_{\text{н}} = 1. \tag{10.18}$$

Usually we are interested in value P_p, characterizing the carrying capacity of the system. In order to find it we initially determine the probability of passing the reconnoitered signal $P_{\text{н}}$. The probability of $P_{\text{н}}$ can be determined as the ratio of average number of signals departing from the line to the average number of signals entering the system in a unit of time. In order to determine the indicated mathematical expectations, it is necessary to solve the system of equations (10.16) and (10.17) relative to probabilities $P_0, P_1, \ldots, P_n, \ldots, P_{n+k}, \ldots$

We will apply the following method for solving the system of equations. With the help of system of equations (10.16) we express all $P_j (j = 1, \ldots, n + k)$ through P_0, after which we substitute their values in (10.17) and find P_0. Further, knowing P_0, we determine P_j.

From the first equation of system (10.16) we find

$$P_1 = \frac{\lambda}{\mu} P_0.$$

From the second equation

$$P_2 = \frac{\lambda^2}{2\mu^2} P_0,$$

from i equation $(i \leq n)$

$$P_i = \frac{\lambda^i}{i! \mu^i} P_0. \tag{10.19}$$

For $i = n + 1$ with the help of the corresponding equation we find

$$P_{n+1} = \frac{\lambda^{n+1}}{n! \mu^n (n\mu + \chi)} P_0.$$

In a general case we have

$$P_{n+k}=\frac{\lambda^{n+k}}{n!\,\mu^n\prod_{j=1}^{k}(n\mu+j\chi)}P_0. \qquad (10.20)$$

Substituting the resulting values for P_i and P_{n+k} in (10.17), we obtain

$$\sum_{i=0}^{n}P_i+\sum_{k=1}^{\infty}P_{n+k}=1. \qquad (10.21)$$

Here

$$\sum_{i=0}^{n}P_i=P_0\sum_{i=0}^{n}\frac{\lambda^i}{i!\,\mu^i},$$

$$\sum_{k=1}^{\infty}P_{n+k}=P_0\sum_{m=1}^{\infty}\frac{\lambda^{n+m}}{n!\,\mu^n\prod_{j=1}^{m}(n\mu+j\chi)}.$$

Hence

$$P_0=\frac{1}{\sum_{i=0}^{n}\frac{\lambda^i}{i!\mu^i}+\sum_{m=1}^{\infty}\frac{\lambda^{n+m}}{n!\mu^n\prod_{j=1}^{m}(n\mu+j\chi)}}. \qquad (10.22)$$

Further, in accordance with the accepted method with the help of (10.19), (10.20), and (10.22) we find

$$P_i=\frac{\lambda}{i!\mu^i}\,\frac{1}{\sum_{j=0}^{n}\frac{\lambda^i}{i!\mu^i}+\sum_{m=1}^{\infty}\frac{\lambda^{n+m}}{n!\mu^n\prod_{j=1}^{m}(n\mu+i\chi)}}, \qquad (10.23)$$

$$P_{n+k}=\frac{\lambda^{n+k}}{n!\mu^n\prod_{j=1}^{k}(n\mu+j\chi)}\,\frac{1}{\sum_{i=0}^{n}\frac{\lambda^i}{i!\mu^i}+\sum_{m=0}^{\infty}\frac{\lambda^{n+m}}{n!\mu^n\prod_{j=1}^{m}(n\mu+j\chi)}}. \qquad (10.24)$$

Let us turn to the determination of what we are interested in, the mathematical expectation $\overline{m}_k$ of signals which are found in line

$$\bar{m}_k = \sum_{k=1}^{\infty} kP_{n+k}, \qquad (10.25)$$

where P_{n+k} is determined with the help of (10.24).

The average number of signals leaving the line in a unit of time will be determined as the ratio $\bar{m}_k$ to the average time of waiting in line $t_{\text{ож}}$.

The unknown probability of omission of reconnoitered signal $P_{\text{н}}$ in accordance with the determination is equal to the ratio of average number of signals leaving the line to the average intensity of flow of signals λ, i.e.,

$$P_{\text{н}} = \bar{m}_k \frac{\chi}{\lambda}. \qquad (10.26)$$

Accordingly the carrying capacity of the electronic reconnaissance system will be determined by the formula

$$q = 1 - \frac{\chi}{\lambda} \sum_{k=1}^{\infty} kP_{n+k}. \qquad (10.27)$$

It is natural that the carrying capacity of a system with waiting is higher than a system with rejections. As waiting time decreases $(\chi \longrightarrow \infty)$ the system with waiting approaches the system with rejections.

In the queueing theory it is proven that in systems with unlimited waiting time $(\chi \longrightarrow 0)$ there is not always a stationary regimen, i.e., it is not always possible to use the system of algebraic equations (10.16) instead of differential equations (10.15).

A stationary regimen exists if the average number of requests, arriving during a period of time equal to the average time of handling one request, does not exceed number of channels of the queueing system, i.e., if

$$\frac{\lambda}{\mu} < n.$$

In many cases the product n_μ is hundreds of times greater than χ, which makes it possible to significantly simplify formulas (10.22), (10.23), and (10.24), and namely at $\frac{\lambda}{\mu} < n$ to obtain

$$\begin{gathered}
P_0 = \frac{1}{\sum\limits_{i=0}^{n} \frac{\lambda^i}{i!\mu^i} + \frac{\lambda^{n+1}}{n!\left(n - \frac{\lambda}{\mu}\right)\mu^{n+1}}}, \\
P_i = \frac{\lambda^i}{i\mu^i} P_0 \quad (0 < i < n), \\
P_{n+k} = \frac{\lambda^{n+k}}{n!\,\mu^{n+k} n^k} P_0 \quad (k \geqslant 1), \\
\bar{m}_k = \frac{\frac{\lambda^{n+1}}{nn!\left(1 - \frac{\lambda}{n\mu}\right)^2 \mu^{n+1}}}{\sum\limits_{k=0}^{n} \frac{\lambda^k}{k!\mu^k} + \frac{\lambda^{n+1}}{n!\left(n - \frac{\lambda}{\mu}\right)\mu^{n+1}}}.
\end{gathered} \qquad (10.28)$$

Example. Let us determine the probability of passing the signal $P_н$ by a two-channel receiver, carrying out electronic reconnaissance of a field of detection and guidance radar, in which the average intensity of target sweep comprises λ = 10 sweeps/s. Average time of continuous operation of radar $1/\chi$ = 10 minutes. Average time for treatment of signal by reconnaissance system comprises $\frac{1}{\mu} = 10^{-3}$s. In this case $\frac{\lambda}{\mu} = 10^{-2} \ll n$, i.e., in the system there is the possibility of a stationary regimen. Furthermore, $n\mu = 2 \cdot 10^3 \geq \chi = 1/600$, therefore it is possible to use formulas (10.28).

In accordance with (10.28):

$$\begin{gathered}
P_0 = 0{,}99, \\
P_1 = 10^{-2} \cdot 0{,}99 = 0{,}0099, \\
P_2 = \frac{1}{2} 10^{-4} \cdot 0{,}99 \approx 0, \\
P_н \approx 0.
\end{gathered}$$

Thus, practically in this case the system of reconnaissance will be free with a probability of 0.99, and the probability of passing a signal is close to zero.

The resulting equations and formulas also make it possible to solve many other problems of electronic reconnaissance, both directly connected with combat application and also with planning of the corresponding equipment.

10.3. Block Diagram of an Electronic Reconnaissance Station

A typical electronic reconnaissance station consists of an antenna unit, receiver, analyzer of parameters of incoming signal, direction finding unit, a unit for memory and treatment of information received, a telemetric device, control equipment, and a power unit. A block diagram of an electronic reconnaissance station is shown in Fig. 10.2.

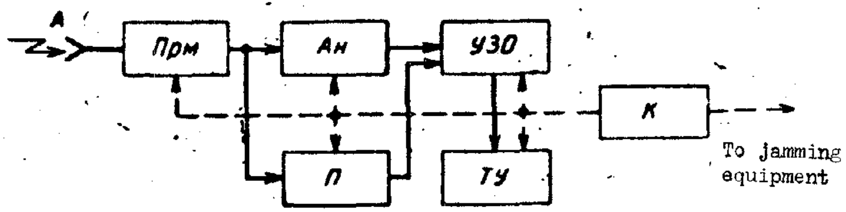


Fig. 10.2. Block diagram of electronic reconnaissance station [See Designation List].

The antenna unit (A) should be broad-band, possess a high carrying capacity, and ensure direction finding of the source with the necessary accuracy. Furthermore, the antenna for an electronic reconnaissance station should have minimum side lobes and a good high-frequency bypassing capacity of fields generated by the transmitting antennas mounted on aircraft, otherwise there is the possibility of a false determination of direction at the direction finding source. Frequently it is impossible to satisfy all requirements with the help of one antenna, therefore usually several antennas are used which cover the entire reconnoitered frequency range. Sometimes a special high-directional antenna is used for finding the direction of reconnoitered devices.

Receivers at electronic reconnaissance stations (PRM) are characterized by the following basic parameters:

- overlapping frequency range;
- time for retuning (carrying capacity);
- sensitivity;
- accuracy in determination of parameters of signals received;
- bearing discrimination;
- method of searching the reconnoitered signal based on carrier frequency and probability of its detection.

The most important technical characteristic of a reconnaissance receiver is full frequency range, in which with its help it is possible to carry out the search of the reconnoitered signals. It is desirable, that one reconnaissance receiver cover as wide a frequency range as possible in which the most important radio electronic devices of the enemy can operate.

The analyzer (An) of parameters of incoming signals serves for an appraisal of parameters and identification of form of the reconnoitered electronic device.

With its help, for example, it is possible to measure time, spectral, and power parameters of incoming signals, and also to make a determination of emission polarization of the reconnoitered device. Time parameters of signals include:

- duration of signals and time intervals between them;
- form of modulating function.

Spectral parameters of signals include: high-frequency spectrum and spectrum of signal envelope. The power characteristic of the incoming signal is its function of spectral density. Analyzers are characterized by number of parameters measured, range of measurements, accuracy, and bearing discrimination.

The direction finder device (П) serves for the determination of angle of arrival of radio waves and, consequently, determination of position of the reconnoitered device. Direction finders have high requirements based on the following parameters:

- high speed operation (within the limit of possibility of measurement of bearing on one pulse);
- accuracy of direction finding;
- bearing discrimination.

The unit for storage and processing of information (UZO) ensures automatic memorization of parameters of signals received: frequency, duration of pulses, period of following, etc. This unit, on the basis of data issued by analyzer, should make the identification of form of reconnoitered device. Identification of form is frequently made by the operator of reconnaissance station. In principle the automatic identification of form is possible with the help of electronic digital computers.

From the point of view of jamming the most important characteristic of the store-process unit is the accuracy and duration of memorization of the carrier frequency. For this purpose, at present quite a few different devices have been developed; some of them will be examined below.

Many parameters of incoming signals can be stored by means of recording them on magnetic tape with the help of video tape recordings or by means of photographing the indicator screens. Results of electronic reconnaissance can also be recorded in the storage unit of the computer.

The telemetric device (TU) serves for transmission of reconnaissance information to the ground. Of particular value are telemetric devices in the carrying out of preliminary electronic reconnaissance with help of artificial earth satellites and pilotless reconnaissance planes. At electronic reconnaissance stations, ensuring directly the means of jamming, telemetric devices can be absent, since reconnaissance information in this case is used directly in the process of overcoming the enemy PVO for the organization of jamming.

Control equipment (K) ensures the automatic or semiautomatic control of the operation of separate units. With its help control of the reconnaissance station on the whole is carried out. An important function of control equipment is delivery of necessary signals to countermeasure equipment.

Determination and storage of carrier frequency of the reconnoitered electronic device are one of the most important functions of the electronic reconnaissance station. The methods of determination and storage of frequency which are used in electronic reconnaissance are specific. The specificity of methods of determination and storage of carrier frequency is caused, on the one hand, by a limitation of reconnaisance time and, on the other hand, by the wide range of reconnoitered frequencies.

At present two basic methods are used for determination of frequency: nonscanning and scanning.

The nonscanning method makes it possible in principle to determine carrier frequency practically instantly, while scanning methods of determination of frequency require a certain amount of time in connection with the necessity for retuning the receiver. This method of determination of frequency makes it possible to significantly reduce the time for reconnaissance, however, reduction of reconnaissance time yields to impairment of accuracy and bearing discrimination of measurements, or to an increase of volume of equipment.

Scanning methods, on the contrary, with a considerable time of reconnaissance makes it possible to measure carrier frequency with great accuracy and ensure a high degree of bearing discrimination.

10.4. Scanning Methods for Determination of Frequency

The scanning method of determination of frequency is usually realized in a so-called panoramic receiver, the block diagram of which is shown in Fig. 10.3.

A panoramic receiver in the simplest case constitutes a superheterodyne which is retuned automatically or manually in the band of reconnoitered frequencies.

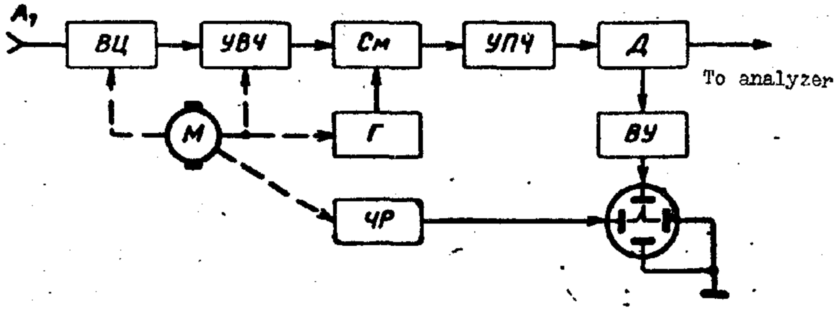


Fig. 10.3. Block diagram of a panoramic receiver [See Designation List].

In the process of frequency search the retuning of the receiver is carried out with help of an electrical motor, which by a definite law simultaneously changes the tuning of the input circuit, high-frequency amplifier, and heterodyne. Simultaneously the motor controls the device for formation of frequency scanning on the screen of the electron-beam tube.

After amplification in the i-f amplifier, detection in the detector, and additional amplification in the video amplifier, the incoming signal is fed to the vertical plates of the indicator, as a result of which a pulse is formed on the screen. Its position on the frequency scanner determines the carrier frequency of the reconnoitered device.

An important characteristic of the panoramic receiver is the time for scanning the carrier frequency (time of reconnaissance).

Usually the scanning of the entire operational frequency range is carried out periodically with a period T_{Π} based on the sawtooth law (Figs. 10.4 and 10.5). Therefore, during reconnaissance of carrier frequency of a continuous signal maximum time of scanning does not exceed T_{Π}. More complicated is the determination of carrier frequency of short-term operational signals. A graphic presentation of this is given by the frequency-time diagram of frequency scanning which is depicted in Fig. 10.4. As can be seen from the figure, continuous signal (f_{H}) is detected with a probability equal to a unit, whereas detection (and, consequently, measurement of frequency) of pulse signal is not always possible. In a general case the process

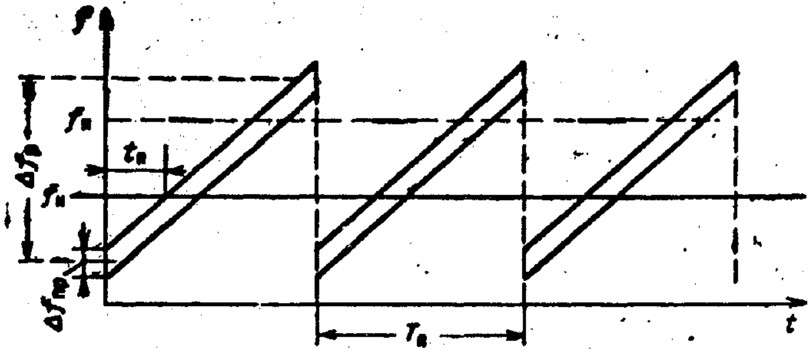


Fig. 10.4. Frequency-time diagram, illustrating the scanning method of frequency determination.

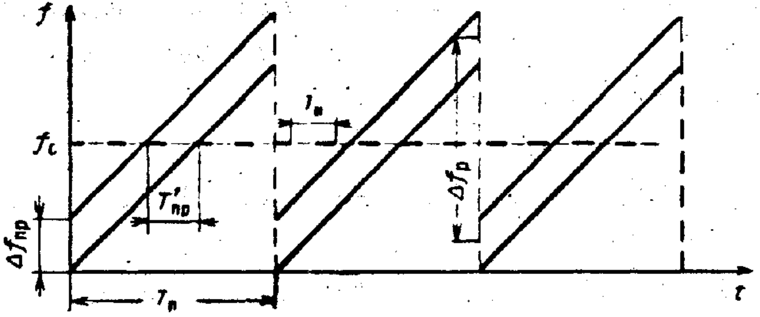


Fig. 10.5. Frequency-time diagram, explaining a slow scanning of frequency.

of detection and measurement of frequency of pulse signal bears a probabilistic nature. Depending on the ratio of the period for retuning and the duration of the signal of the reconnoitered device three scanning methods of frequency determination are distinguished:

- slow scan,
- fast scan,
- scan with moderate speed.

Slow Scan

During slow scanning the time for retuning of the receiver $T'_{\text{пр}}$ on the breadth of its transmission band is greater than the period of following pulses $T_{\text{и}}$ (Fig. 10.5), i.e.,

$$T'_{\text{пр}} > T_{\text{и}}.$$

If the determination of frequency can be done on one pulse, then slow scanning ensures the probability of detection of the periodic pulse signal P = 1 during the time for retuning $T_{\text{п}}$. A serious deficiency of slow scanning is the great amount of time for handling, small carrying capacity, and accordingly a small probability of reconnaissance of electronic means which operate momentarily.

For reducing the time of reconnaissance in an assigned range and the time for retuning it is necessary to expand the width of the transmission band of the receiver. Therefore panoramic receivers with slow scanning, as a rule, are broadband. Bandwidth of such receivers is approximately equal to

$$\Delta f_{\text{пр}} = (0,1 \div 0,01)\,\Delta f_{\text{г}},$$

where $\Delta f_{\text{р}}$ — range of retuning (range of reconnoitered frequencies).

Accuracy of determination of carrier frequency with help of such receivers is not great. It is approximately half of the transmission band of the receiver, i.e.,

$$(\delta f)_{\text{макс}} = 0,5\Delta f_{\text{пр}} = (0,05 \div 0,005)\,\Delta f_{\text{р}}.$$

Sensitivity of receiving devices with slow scanning, due to considerable passband, cannot be high. Frequently these receivers are carried out on a circuit of straight amplification with reconstructed input circuits.

Time of guaranteed detection during slow scanning is determined by the period of retuning $t_{\text{го}} = T_{\text{п}}$.

Fast Scanning

During fast scanning the time for retuning of the receiver in the entire operational range $(\Delta f_{\text{р}})$ is less than the duration of the incoming signal (Fig. 10.6), i.e., $T_{\text{п}} < \tau_{\text{и}}$.

Retuning time in this case is extraordinarily great (hundreds and thousands of megacycles per second in microseconds). Such

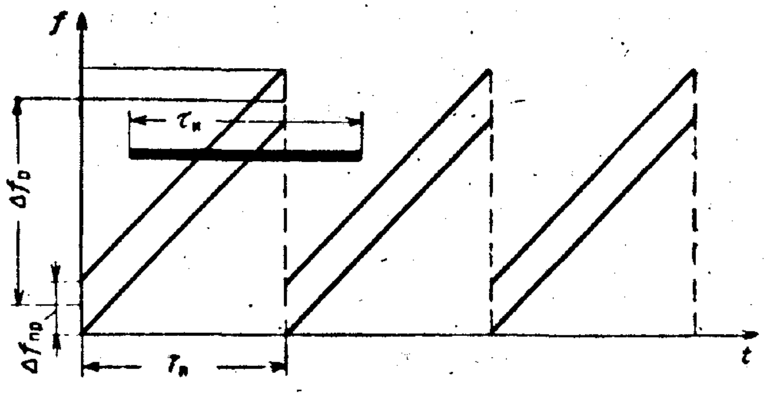


Fig. 10.6. Frequency-time diagram, characteristic for fast scanning of frequency.

speeds can be ensured only by electronic methods [91, 92].

Retuning rate cannot be infinitely great. It is limited by acceptable limits of lowering of sensitivity, accuracy, and resolving power during a determination of frequency, taking place due to the inertness of resonance devices.

Resonance devices, being under the influence of signals with variable frequency, are characterized by a dynamic frequency-response curve, by which is understood the dependence of the ratio of output voltage to input from detuning of the relatively inherent resonance frequency of the system at a fixed rate of retuning.

Dynamic characteristics depend both on the parameters of the resonance system (for example, width of static passband) and also on the speed of retuning or speed of frequency shift of the external signal. In Fig. 10.7, for an illustration, is depicted a family of frequency-response curves of a single oscillatory circuit [85]. The parameter of the family is the coefficient ξ, equal to

$$\xi = \sqrt{\frac{\pi}{\gamma}\Delta f_{пр}},$$

where $\gamma = \frac{df}{dt}$ — rate of frequency shift of actuating voltage (speed of retuning); $\Delta f_{пр}$ — width of static characteristics of oscillatory circuit on a level of 0.707.

From an analysis of the cited frequency-response curves the following conclusions can be made:

– with an increase of retuning rate the maximum of characteristics shifts in the direction of frequency shift (in this case in the

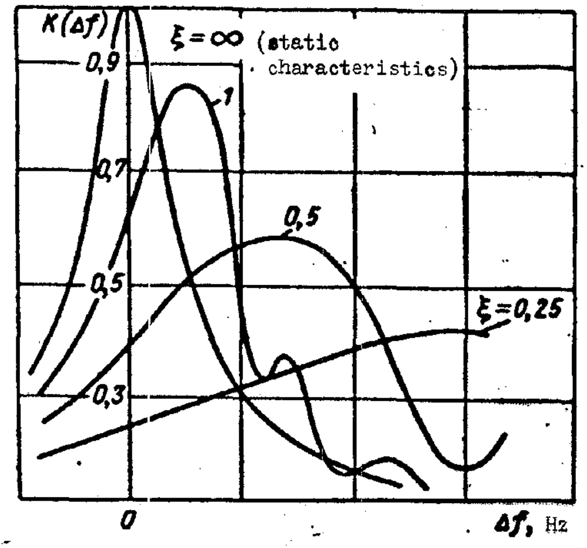


Fig. 10.7. Dynamic frequency-response curves of a single oscillatory circuit.

direction of an increase), and the value of output voltage decreases;

- transmission bandwidth on a level of 0.707 is also increased with an increase of retuning rate;
- additional maxima of frequency-response curves appear.

The enumerated peculiarities are the reason for impairment of characteristics of a reconnaissance receiver of this class, they:

- decrease the sensitivity of the receiver;
- lower the accuracy and resolving power;
- limit the speed of retuning and, consequently, the time of reconnaissance;
- distort the parameters of the reconnoitered signal (form, duration).

For panoramic receivers with rapid retuning there is an interconnection between passband of the resonance system and retuning rate; an increase of retuning rate leads to loss of accuracy of measurement of carrier frequency and a lowering of sensitivity. Actually, the optimum passband $\Delta f_{\text{пр}}$ of a radio receiver and the pulse width τ, forming on the output as a result of rapid retuning, in the case of approximation of form of pulse and frequency-response curve of the receiver by rectangles are connected in the first approximation by the following relationship:

$$\Delta f_{\text{пр}} = \frac{1}{\tau}.$$

Pulse width at the assigned retuning rates γ and passband $\Delta f_{\text{пр}}$ is equal to $\tau=\frac{\Delta f_{\text{пр}}}{\gamma}$. It follows from this that

$$\Delta f_{\text{пр}}\approx\sqrt{\gamma}.$$

More exact investigations [85] show that in case of a bell-shaped frequency-response curve

$$\Delta f_{\text{пр}}\approx\frac{1}{\sqrt{\pi}}\sqrt{\gamma}.$$

Thus, each rate of retuning has its own optimum band. By reducing the time of scanning we lose out in accuracy of determination of frequency and, conversely, increasing the accuracy of determination of frequency, simultaneously should increase the time of reconnaissance.

Example. If $\Delta f_{\text{пр}}=10$ MHz, then the maximum speed of retuning allowed is $\gamma_{\text{макс}}=\pi\cdot10^{8}$ [MHz/s].

Loss of sensitivity depending on rate of scanning on frequency γ can be evaluated with the help of expression [23]:

$$a=\left[1+0{,}195\left(\frac{\Delta f_{\text{р}}}{T\Delta f_{\text{пр}}^{2}}\right)^{2}\right]^{-\frac{1}{4}}=$$
$$=\left[1+0{,}195\left(\frac{\gamma}{\Delta f_{\text{пр}}^{2}}\right)^{2}\right]^{-\frac{1}{4}},$$

where a — loss of sensitivity relative to the receiver with a zero rate of scanning for frequency (in decibels); $\Delta f_{\text{р}}$ — range of reconnoitered frequencies; T — period of frequency scanning; $\Delta f_{\text{пр}}$ — passband of receiver; $\gamma=\frac{\Delta f_{\text{р}}}{T}$ — rate of scanning.

For reducing the dynamic effect it is necessary at a constant rate of retuning γ to increase the passband of the resonance system, but this, in its turn, leads to a decrease of sensitivity of the receiver and accuracy of measurements.

The simultaneous guarantee of a considerable rate of retuning and high resolving power for frequency can be successfully achieved in the receiver with a compression of pulses [87, 88]. Here, in point of fact, is used the same principle of an increase of resolving power as is used in wideband radar with coding.

In Fig. 10.8 is depicted a time diagram of pulses at the output of the i-f amplifiers explaining the possibility of improvement of resolving power for frequency in a receiver with compression of pulses.

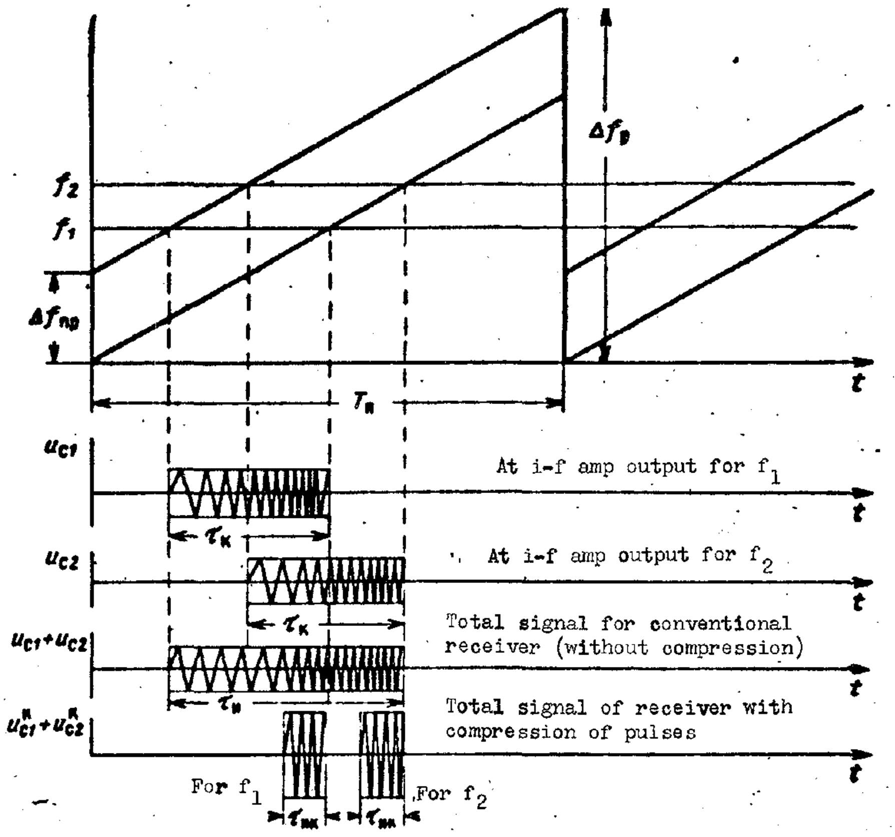


Fig. 10.8. Time diagrams, explaining the possibility of improvement of resolving power for frequency in a reconnaissance receiver with compression of pulses.

If two continuous signals with different frequencies f_1 and f_2, are influencing the reconnaissance receiver, then as a result of retuning the heterodyne at the output of the i-f amplifier with pass-band $\Delta f_{пр}$ frequency-modulated pulses with a duration $\tau_{к}$ will be formed. In the event of summation of these pulses in a conventional panoramic receiver one pulse with a duration $\tau_{н}$, will be formed, and there is no possibility to solve the incoming signals for frequency.

In a receiver with compression of pulses the signals from the output of the i-f amplifier enter the dispersion filter, in which compression of signals to durations of $\tau_{нк}$ takes place. As a result the signals are solved for duration, and, consequently, for frequency.

Thus the resolving power for frequency is increased. It has been determined that this increase is proportional to the square root from the coefficient of compression (compression). For example, a receiver with compression of pulses, which has been readjusted in a frequency range with a rate of γ = 100 MHz/μm·s and possesses a coefficient of compression of 100, has a resolving power for frequency equal to 1 MHz, i.e., 10 times higher than the resolving power of typical panoramic receivers which have the same rate of retuning.

A block diagram of a receiver with compression of pulses is shown in Fig. 10.9.

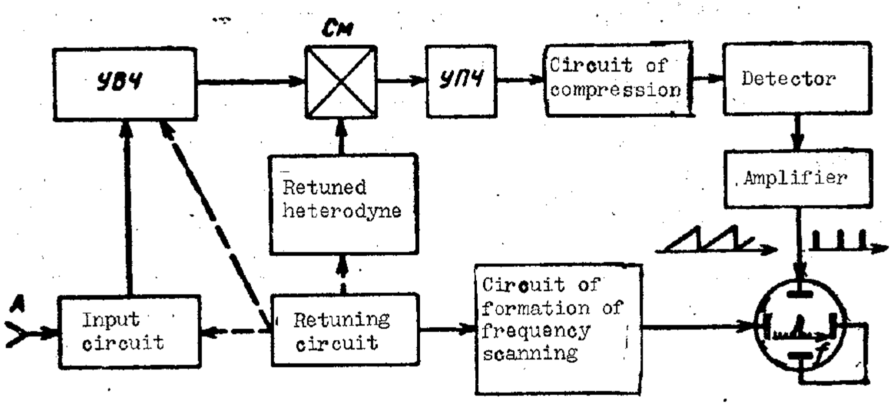


Fig. 10.9. Block diagram of reconnaissance receiver with compression of pulses.

The actuating signal is amplified by the wideband r-f amplifier and enters the mixer, where voltage of the heterodyne, which is periodically retuned for frequency, also arrives. Frequency shifting of the heterodyne is carried out with the help of a retuning circuit, which also controls the circuit of formation of frequency scanning.

As a result of conversion of signals in the mixer, at the input of the i-f amplifier signals with a frequency which is linearly variable in time are obtained.

For conversion of pulses with a linearly variable frequency into signals with less duration and greater amplitude, in the receiver a circuit of compression (dispersion filter) is used which constitutes a high-frequency line of delay with taps, in each of which is a band filter. After circuit of compression the signal is detected, amplified, and fed to the vertical deflection plates of the electron-beam tube. Frequency of signals is determined by the position of pulse on the frequency scanner.

As was already noted above, rapid scanning leads to an impairment of characteristics of the reconnaissance receiver. Realization of rapid scanning requires considerable complication of equipment.

Besides slow and fast, it is possible to use probabilistic scanning (scanning with average speed) thus ensuring the best conditions for a compromise between speed of retuning and accuracy of determination of frequency at an assigned probability of electronic reconnaissance.

Scanning with Moderate Speed

This type of scanning for frequency is most characteristic for electronic reconnaissance. Time for retuning the reconnaissance receiver T'_{np} for the width of its passband during scanning with moderate speed is determined by the following relationship:

$$kT_c > T'_{\text{np}} > \tau_c.$$

where T_c – period of tracking of pulses; τ_c – duration of reconnoitered pulses; k = 1, 2, 3.

A distinctive peculiarity of scanning with moderate speed is the lack of guaranteed detection of the operation of pulse radar during one period of retuning of the reconnaissance receiver. In other words, the probability of detection of the reconnoitered signal, in the examined case, in principle is always less than a unit $(P_{обн}<1)$. For this reason scanning with moderate speed is sometimes called probabilistic scanning.

An analysis of scanning with moderate speed can be conveniently conducted with the help of the theory of random pulse flows which was developed by N. M. Sedyakin [89, 90]. In the examined case there are two flows of pulses (Fig. 10.10). The first characterizes the flow of pulses of the reconnoitered device with duration τ_c and period of tracking T_c. The second characterizes the readiness of the reconnaissance receiver to handle the flow of signals; the parameters of this flow are period of retuning $T_п$ and time of retuning the receiver $T'_{пр}$ on a value equal to the passband.

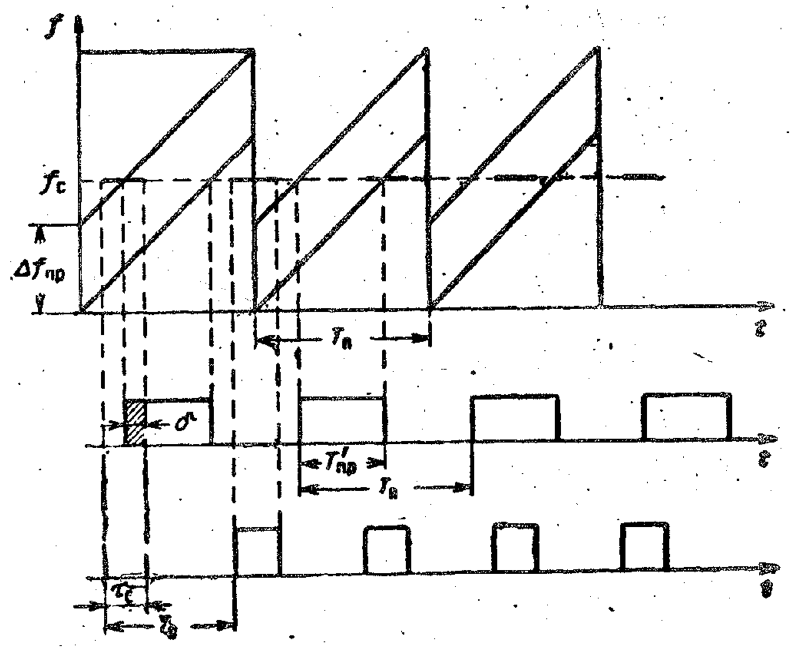


Fig. 10.10. Frequency-time diagrams, illustrating frequency scanning with moderate speed.

Detection occurs at the moments of "engagement" of flows. If the duration of "engagement" δ is sufficient for reliable operation of the reconnaissance receiver, then simultaneous with detection it is also possible to determine the frequency of the reconnoitered device.

The theory of random pulse flows gives the following formulas for the mean frequency of tracking F and mathematical expectation of duration of pulses of flow of coincidences $\bar{\tau}_\delta$;

$$F(\delta)=\frac{\tau_c+T'_{\text{пр}}-2\delta}{T_c\cdot T_{\text{п}}},$$

$$\bar{\tau}_\delta=\frac{\tau_c T'_{\text{пр}}}{\tau_c+T'_{\text{пр}}}.$$

Here $F(\delta)$ — mean frequency of tracking of pulses on the output of the reconnaissance receiver, and the duration of which is not less than δ.

Probability of "engagement" of independent flows for duration δ during one period of tracking of pulses T_c is determined by the formula

$$P_{\text{з}}=\frac{\tau_c+T'_{\text{пр}}-2\delta}{T_c}.$$

If δ – duration of minimum pulse necessary for realization of reconnaissance, then P_3 determines the probability of detection of radar in one period T_c. Considering that

$$T'_{\text{пр}}=\frac{\Delta f_{\text{пр}}}{\gamma}=\frac{\Delta f_{\text{пр}}}{\Delta f_{\text{р}}}T_{\text{п}},$$

for the probability of detection of radar during the time of retuning the receiver we obtain

$$P_{\text{обн}}(T)=\frac{\tau_c+\dfrac{\Delta f_{\text{пр}}}{\Delta f_{\text{р}}}T_{\text{п}}-2\delta}{T_c}.$$

If it is considered that $\tau_c\ll T'_{\text{пр}}$, $\delta\approx 0$, then the formula is simplified

$$P_{\text{обн}}(T_c)=\frac{\Delta f_{\text{пр}}}{\Delta f_{\text{р}}}\frac{T_{\text{п}}}{T_c}.$$

Probability of detection of the signal during the time $t_p > T_c$ can be estimated by the expression

$$P = 1 - e^{-\frac{\Delta f_{\text{пр}}}{\Delta f_{\text{р}}} \frac{T_{\text{и}}}{T_{\text{с}}}}.$$

10.5. Nonscanning Methods for Determination of Frequency

The essence of a nonscanning method of frequency determination lies in the fact that reconnaissance is conducted simultaneously in all sectors of operational frequency bands.

Receiving devices, using nonscanning methods of frequency determination, ensure simultaneous reception in a wide band of operating frequencies without retuning of heterodynes or filters. Time for reconnaissance of frequency with nonscanning methods can be very short, since all components of the spectrum of the received signal are exposed simultaneously and practically instantly. At present the following nonscanning methods of frequency determination are known:

- reconnaissance with the application of frequency discriminators;
- functional (interference) methods;
- intelligence with the help of multichannel receivers.

Reconnaissance of Frequency with the Help of Frequency Discriminators

The possibility of frequency determination with the help of frequency discriminators is based on the capacity of the latter to convert deflections of frequency from a rated value into voltage which is proportional to this deflection.

The simplest devices for determination of frequency can be ordinary frequency discriminators, which were examined, for example, in [36]. However, for purposes of intelligence it is better to use a somewhat different arrangement of frequency discriminators. This is connected with the peculiarities of coupling them with receiving and indicator devices at the reconnaissance station.

In Fig. 10.11 is depicted a block diagram of a reconnaissance receiver with frequency discriminator [87]. After amplification in the wideband amplifier , the accepted signal, moves to the discriminator, the output signal of which, after corresponding amplification in the amplifiers, moves to the horizontal and vertical deflecting plates of the electron-beam tube.

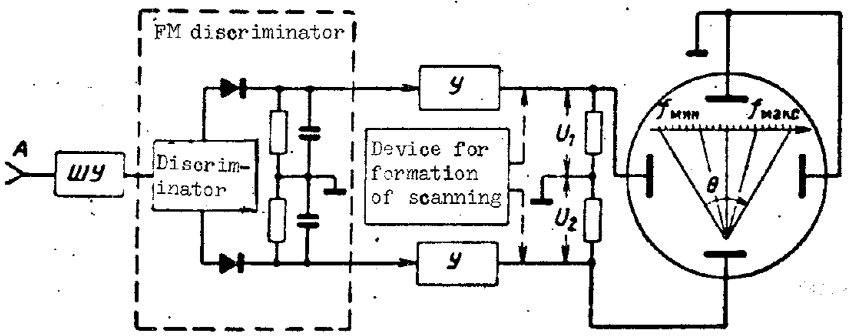


Fig. 10.11. Reconnaissance receiver with frequency discriminator.

Voltage U_1 is fed to the horizontal deflecting plates, and U_2 – to the vertical deflecting plates of the indicator. The luminous dot on the indicator screen will be deflected with a frequency shift f of the input signal. If the signal on frequency $f_{мин}$ produces the deflection shown in Fig. 10.11, then the signal with $f_{макс}=2f_{мин}$ creates on the screen a spot which is displaced relative to initial by angle θ. By modulating the corresponding scanning voltages with voltages U_1 and U_2, it is possible to create a frequency scan. Frequency of the incoming signal determines the value of angle deflection θ – the trace of the electron spot (scanning trace) relative to the beginning of the reading.

Unmodulated harmonic oscillation is observed on the indicator screen in the form of one line. A frequency modulated signal is observed on screen in the form of two lines, the angular distance between which depends on deviation of frequency.

This receiver ensures determination of frequency of a large number of reconnoitered electronic devices under the condition that signals received from them do not coincide in time. Signals coinciding will elicit on the screen a line, the position of which will be determined by the vector sum of signals received. However, with help of spatial selection in a number of cases it is possible to avoid simultaneous reception of two or several signals with different frequency.

The frequency discriminator of the receiver is made up of passive elements constituting segments of long lines or waveguide sections. The set of sections covers the decimetric and centimetric ranges of waves.

Figure 10.12 shows the circuit of a passive frequency discriminator, fulfilled on segments of long lines.

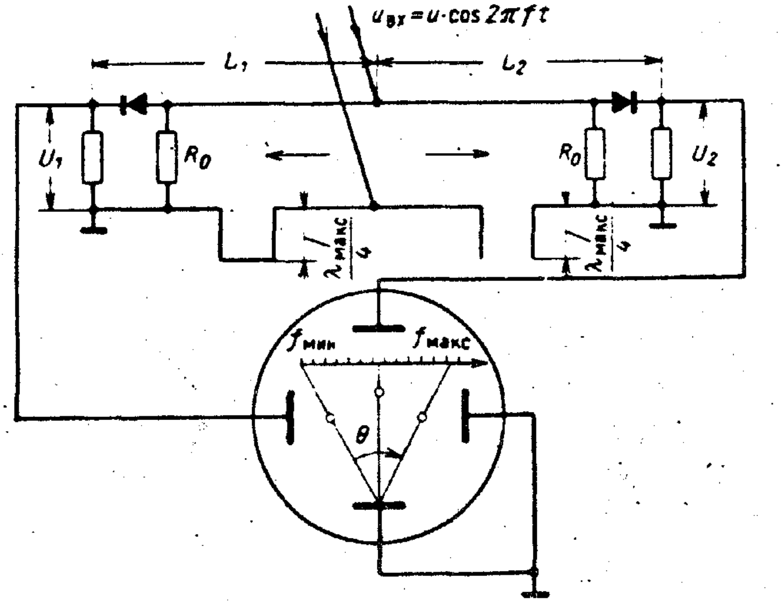

$u_{вх} = u \cdot \cos 2\pi f t$
L_1
L_2
U_1
R_0
R_0
U_2
$\frac{\lambda_{макс}}{4}$
$\frac{\lambda_{макс}}{4}$
$f_{мин}$
$f_{макс}$
θ

Fig. 10.12. Circuit of passive frequency discriminator, carried out on segments of long lines.

The signal of the reconnoitered device enters the middle of line and spreads along this line to various sides (to the left and to the right). The right and left arms of the lines are equal in length ($L_1 = L_2$) and have identical wave impedances w and load impedances $R_0 = w$. Besides the segments of the line have accordingly opened and closed sticks. Length of loops is selected as equal to a fourth of the maximum measured wavelength $\lambda_{макс}$ (corresponding to least measured frequency $f_{мин}$).

Considering that input impedances of the quarter-wave short-circuited and opened stubs are determined in this case accordingly by formulas

$$z_{кз} = w \operatorname{tg} \frac{2\pi}{\lambda} \frac{\lambda_{макс}}{4},$$

$$z_{1\cdot 3} = -w \operatorname{ctg} \frac{2\pi}{\lambda} \lambda_{макс},$$

for voltages U_1 and U_2 it is possible to write

$$U_1 = k_1 u \operatorname{tg} \frac{\pi}{2} \frac{\lambda_{макс}}{\lambda} \cos(\ t + \varphi_1),$$

$$U_2 = -k_2 u \operatorname{ctg} \frac{\pi}{2} \frac{\lambda_{макс}}{\lambda} \cos(\omega t + \varphi_2).$$

If the right and left arms of the line are identical, then it is possible to consider $k_1 = k_2$, and $\phi_1 = \phi_2$. Then the ratio of voltages

$$K(\lambda) = -\frac{U_1}{U_2} = -\operatorname{tg}^2 \frac{\pi}{2} \frac{\lambda_{макс}}{\lambda}$$

simply determines the frequency of input signal $u_{вх} = u \cos 2\pi f t$. Figure 10.13 represents the dependence of $K(\lambda)$ on the ratio $\lambda_{макс}/\lambda$.

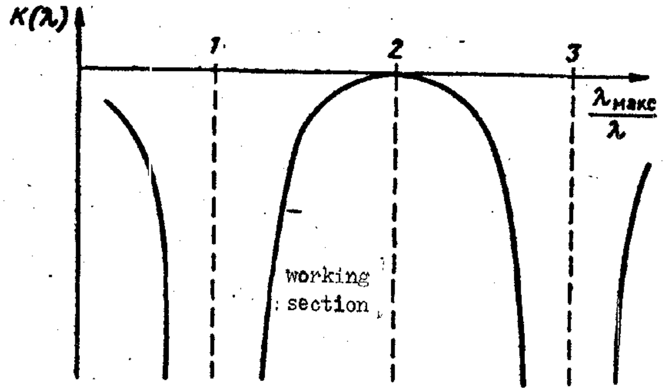


Fig. 10.13. Dependence of standardized output voltage of passive frequency discriminator on wavelength of input signal.

Frequency range, within the limits of which the simple measurement of frequency is possible, is determined by the dimensions of short-circuited and opened stubs. For quarter-wave stubs the ratio $K(f) = U_1/U_2$ will be unambiguous approximately within the limits of one octave. On higher frequencies, where application of a double line is impossible, passive frequency discriminators are used which are carried out on segments of waveguides with the inclusion of stubs in the narrow and wide walls.

Receivers with frequency discriminators are promising. With their help it is possible to determination of frequency in a wide range with a relatively high degree of accuracy.

For example, the experimental American reconnaissance receiver with a WHIP frequency discriminator determines frequency in the range of 500-10,000 MHz with an accuracy of 1-2%. Sensitivity of this receiver is twice as high as the sensitivity of contemporary detector receivers with straight amplification.

Interference Methods of Determination of Frequency

At the basis of the interference method of determination of carrier frequency lies the well-known dependence of shift of phases on length of path and frequency [96]. Principle of construction of reconnaissance receivers, in which the interference method of measurement of frequency is used, can be explained with help of Fig. 10.14.

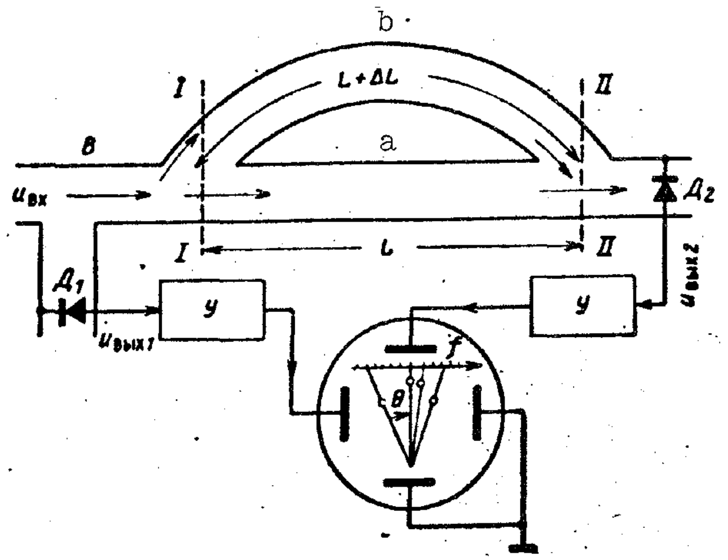


Fig. 10.14. Functional circuit of the interference meter of frequency.

The accepted oscillations are circulated in the common waveguide B. In section I-I the waveguide is branched out and the signal spreads on different waveguides a and b. Length of waveguide b is greater than the length of waveguide a by a certain value ΔL. In section II-II the fields coming from waveguides a and b are geometrically combined. Phases of summable fields will differ by the value

$$\Delta\varphi=\frac{\omega\Delta L}{v_{\Phi}},$$

where v_ϕ – phase rate of propagation of electromagnetic wave in the waveguide.

We will determine voltage on the input of the detector, assuming that the latter is coordinated with the waveguide. In section I-I (before branching) we have

$$u_{\text{вх}}=U_0\cos(\omega t+\varphi_0).$$

On the outputs of waveguides a and b we obtain respectively

$$u_1=\text{к}U\cos\left[\omega\left(t+\frac{L}{v_{\Phi}}\right)+\varphi_0\right],$$
$$u_2=\text{к}U\cos\left[\omega\left(t+\frac{L+\Delta L}{v_{\Phi}}\right)+\varphi_0\right],$$

where L – length of waveguide a: L + ΔL – length of waveguide b; K – constant coefficient.

Resultant voltage will be equal to

$$u_{\text{д вх}}=u_1+u_2=\text{к}2U_0\cos\frac{\omega\Delta L}{2v_{\Phi}}\cos(\omega t+\varphi'_0),$$

where

$$\varphi'_0=\varphi_0+\frac{\omega\Delta L}{2v_{\Phi}}+\frac{\omega_0 L}{v_{\Phi}}.$$

After detection in detector Д_2 we obtain

$$u_{\text{вых}2}=\text{к}_{\text{д}}U\cos\frac{\omega\Delta L}{2v_{\Phi}}, \tag{10.29}$$

where $K_{\text{д}}$ – constant coefficient.

Consequently, output voltage $u_{\text{вых2}}$ is a function of frequency. In Fig. 10.15 is depicted the dependence of output voltage $u_{\text{вых2}}$ on frequency of the actuating signal. Frequency range Δf_p, within the limits of which a simple measurement is possible, is determined by the difference ΔL of lengths of waveguides a and b. From (10.29) it follows that this will be within the limits of any half-wave of the cosine curve, when its argument obtains values from $n\pi$ to $(n + 1)\pi$. Hence the minimum and maximum reconnoitered frequencies are determined by the following equalities:

$$f_{\text{мин}} = \frac{n v_{\Phi}}{\Delta L}, \tag{10.30a}$$

$$f_{\text{макс}} = \frac{(n+1) v_{\Phi}}{\Delta L} \tag{10.30b}$$

where n = 1, 2, 3, ...

The joint solution of (10.30a) and (10.30b) gives

$$f_{\text{макс}} = f_{\text{мин}} \frac{n+1}{n}$$

If n = 1, then a simple determination of frequency is possible within the limits

$$f_{\text{мин}} \leqslant f \leqslant 2 f_{\text{макс}}.$$

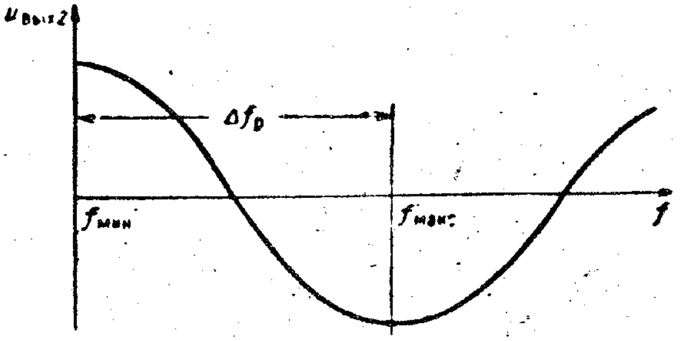


Fig. 10.15. Dependence of output voltage of interference meter on frequency.

Output voltage $u_{\text{вых}2}$ by itself cannot be used directly for measurement of frequency, since its value depends on intensity of signal received. For an exception of this dependence normalization of voltage $u_{\text{вых}2}$ is carried out relative to amplitude of input signal. Input signal is detected in the dector (before branching of waveguide), and in a special computer division of voltage $u_{\text{вых}2}$ into amplitude of input signal $u_{\text{вых}1}$. is carried out. As a result we obtain the ratio

$$K(f)=\frac{u_{\text{вых}2}}{u_{\text{вых}1}}=K\cos\frac{\omega\Delta L}{2v_{\text{ф}}},$$

being only a function of frequency. Such an operation of standardization of output voltage relative to amplitude of input signal can be carried out with help of an electron-beam tube, to the horizontal deflecting plates of which is fed the voltage $u_{\text{вых}1}$, and to the vertical deflecting plates – $u_{\text{вых}2}$. The luminous dot on the screen of the electron-beam tube will be deflected during a frequency shift of input signal. With the help of the corresponding circuits for formation of scanning it is possible to obtain a scan which is convenient for reading of frequency in the form of lines, the angles of inclination of which are proportional to frequency. A brief description of principle of action of such an indicator was given above.

More improved is the interference meter of frequency, using as the basic element a phase detector on a double waveguide tee (Fig. 10.16). The frequency measuring circuit in this case can be made analogous to the circuit of monopulse direction finder with total-difference treatment (Fig. 4.1).

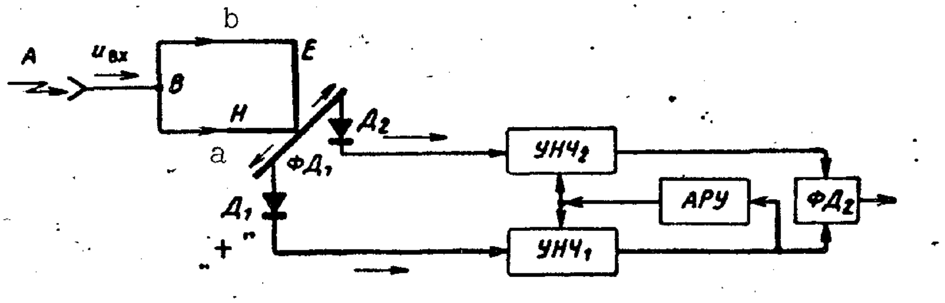


Fig. 10.16. Functional circuit of a device for determination of frequency on a double waveguide tee.

Actuating signal $u_{вх}$ enters arms E and H of the double waveguide tee from antenna A along the waveguide, which branches at point B into two segments a and b of various length. The latter, in their turn, accordingly are connected to arms H and E of the double tee. Electrical lengths of sections a and b differ by a certain value $\Delta \ell_a$, which is equivalent to the difference in geometric lengths of paths on ΔL. On envelope detectors $Д_1$ and $Д_2$, incorporated in the arms of the bridge, will correspondingly be influenced by total and difference field. From the output of the detectors the signals reach the amplifiers of low frequency $UNCh_1$ and $UNCh_2$, after which they move to phase detector FD_2 .

The operation of signal normalization for amplitude is carried out with help of the system of automatic gain control, working from total channel and carrying out the adjustment of amplification in both amplifiers ($UNCh_1$ and $UNCh_2$). Assuming the idealness of functioning of the entire circuit (coordination of detectors with waveguide, identity of amplifiers $UNCh_1$ and $UNCh_2$, idealness of automatic gain control, and so forth.), it is possible in the following way to describe the operation performed by it.

On the output of the antenna a monochromatic signal can be recorded in the following way:

$$u_{вх} = U \cos(\omega t + \varphi_0).$$

In arms H and E we have

$$u_H = U \cos(\omega t + \varphi_0 + \varphi).$$

$$u_E = U \cos(\omega t + \varphi_0).$$

Here

$$\varphi = \frac{\omega \Delta L}{v_\Phi},$$

v_Φ — phase speed of wave in waveguide:

$$v_{\Phi} = \frac{v}{\sqrt{1-\frac{\lambda}{2a}}}.$$

On the input of total and difference channels we obtain

$$u_{\text{с}} = U\left[\cos(\omega t + \varphi_0) + \cos(\omega t + \varphi_0 + \varphi)\right],$$
$$u_{\text{р}} = U\left[\cos(\omega t + \varphi_0) - \cos(\omega t + \varphi_0 + \varphi)\right].$$

On the output of detectors the envelopes

$$u_{\text{д с}} = 2\kappa_{\text{д}} U \cos\frac{\varphi}{2},$$
$$u_{\text{д р}} = 2\kappa_{\text{д}} U \sin\frac{\varphi}{2},$$

where $\kappa_{\text{д}}$ — transmission factor of detectors (identical for $Д_1$ and $Д_2$).

After the amplifier of low frequency

$$u_{\text{с у}} = 2K\kappa_{\text{д}} U \cos\frac{\varphi}{2},$$
$$u_{\text{р у}} = 2K\kappa_{\text{д}} U \sin\frac{\varphi}{2},$$

where K – amplification factor of $\text{UNCh}_{1,2}$.

We select the parameters of the ARU and $\text{UNCh}_{1,2}$ in such a manner that the following equality was ensured,

$$K = \frac{\kappa_0}{2\kappa_{\text{д}} U \cos\frac{\varphi}{2}},$$

where κ_0 – constant coefficient.

Then after multiplication in the second phase detector on its output we obtain

$$u_{\Phi\,\text{д}} = \kappa_{\Phi\,\text{д}}\kappa_0 \operatorname{tg}\frac{\varphi}{2},$$

where $K_{фд}$ — transmission factor of phase detector.

Thus, voltage on the output of the phase detector $u_{фд}$ is a single-valued function of carrier frequency in a sufficiently wide range of waves (Fig. 10.17).

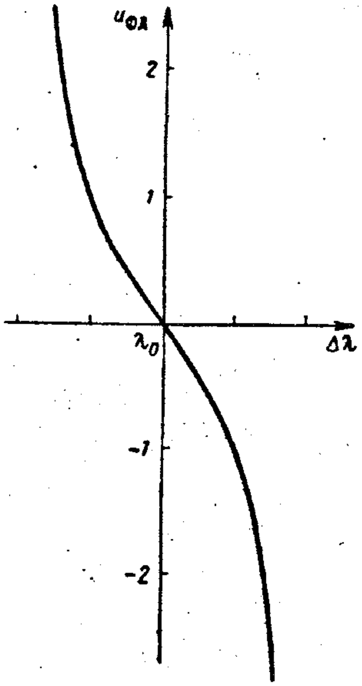


Fig. 10.17. Dependence of voltage on the output of the phase detector on an increase of wavelength.

In principle it would have been possible to obtain the unknown dependence of voltage on frequency directly on the output of the $UNCh_2$. Application of the phase detector FD_2 permits expanding of frequency ranges reconnoitered by the device.

The given consideration confirms the presence of an analogy between the examined interference method of determination of frequency and the monopulse method of direction finding with total-difference treatment. With help of a circuit containing double waveguide bridge it is also possible to develop a system of automatic electron tuning of generator (klystron, backward-wave tubes [LOV]) on the carrier frequency of the accepted signal [96]. In order to ensure the possibility of functioning of interference system of measurement of frequency in the case of the simultaneous arrival of several signals on various frequencies it is necessary to place on the input a dispersion filter, converting the frequency differences of signals into time.

Interference frequency meters possess essential advantages:

- minimum time of reconnaissance of frequency;
- comparatively wide reconnoitered frequency range;
- small volume of equipment.

It is necessary to note the deficiencies of functional meters:

- relatively low sensitivity of receiving devices;
- lowering of accuracy and resolving power during expansion of range of reconnaissance in one device;
- necessity of complication of equipment for determination of frequency of several simultaneously actuating signals.

The low sensitivity of functional meters is explained by the fact that the passband of reconnaissance receiver with a functional meter is very wide. It is equal to the entire reconnoitered frequency range.

Increase of sensitivity of functional meters can be attained at the expense of application of wideband high-frequency amplifiers, the TW tube for example.

Reconnaissance of Carrier Frequency with the Help of Selective Receivers with Straight Amplification

At present it has been determined that basically two types of straight receivers can be used for purposes of electronic reconnaissance:

- single-channel wideband receivers;
- multichannel receivers.

Single-channel wideband receiver. The simplest single-channel wideband straight receiver (aperiodic receiver) consists of an antenna, crystal detector, video amplifier, and indicator (Fig. 10.18). A merit of this receiver is the possibility to completely reproduce

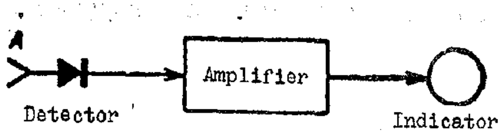


Fig. 10.18. Circuit of single-channel wideband receiver.

information which is included in the signal received. However, its sensitivity is very low and accuracy of measurement of frequency is low. It is determined by approximately half of the width of the transmission band of the antenna or the input filter.

Single-channel wideband straight receivers are used at present only for detection of the actual fact of irradiation.

In contrast to a single-channel receiver a two-channel set has two channels of reception (Fig. 10.19). In each channel there is a resonance filter. In the first channel the resonance circuit is tuned on the lowest, and in second – on the highest frequency of the reconnoitered range (Fig. 10.20). Signals from the outputs of both channels arrive at different groups of deflecting plates of the oscillograph. The angle of deflection of scan line on the indicator screen is a single-valued function of measured frequency. It is natural that the accuracy of measurement of frequency in this case decreases with an increase of reconnaissance range.

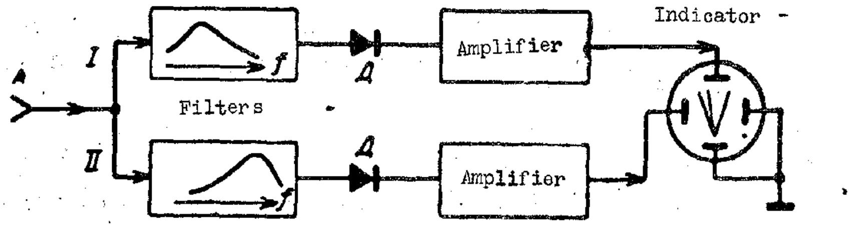


Fig. 10.19. Circuit of a two-channel reconnaissance receiver.

Multichannel receivers. A high degree of accuracy and resolving power can be obtained with the help of multichannel reception. In this case the entire range of reconnoitered frequencies is divided by a system of filters into a number of sub-bands. Bands of transmittance of filters adjoin one another as is shown in Fig. 10.20.

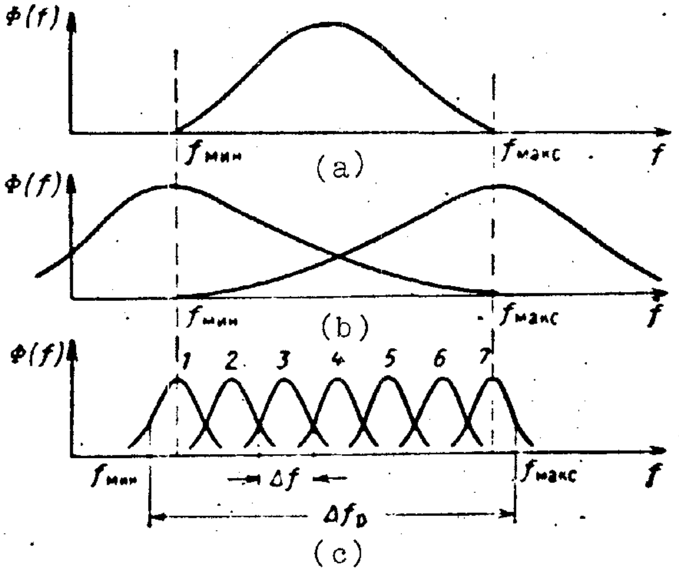


Fig. 10.20. Frequency characteristics of input filters of single-chanel (a) two-channel; (b) and multichannel; (c) receivers.

A block diagram of a multichannel straight receiver with independent channels of reception is presented in Fig. 10.21. Let us remember that a multichannel receiver in this case is not equivalent to a multichannel queueing system.

Bandwidth of transmittance Δf of each filter is selected from the condition of obtaining an assigned accuracy of determination of frequency δf:

$$\Delta f = 2\delta f.$$

The number of filters m depends on assigned accuracy of determination of frequency δf and range of reconnoitered frequencies Δf_p. With identical channels of a receiver

$$m = \frac{\Delta f_p}{2\delta f}.$$

Multichannel receivers are used in stations for preliminary reconnaissance for a rough determination of frequency and identification of form electronic device. The number of channels in them reaches several dozen. Wide use of microminiature blocks, semiconductors, electronic circuits on a solid body, and others indicates the promise of this trend.

In stations of direct electronic reconnaissance use has been

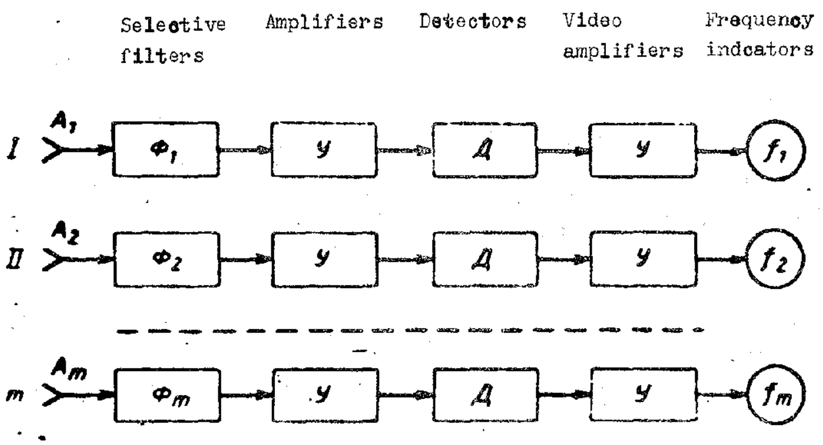


Fig. 10.21. Block diagram of multichannel receiver.

made of a multichannel receiver which guarantees great accuracy with a lesser number of filters [23]. This receiver we will conditionally call matrix. A block diagram of the matrix receiver is shown in Fig. 10.22.

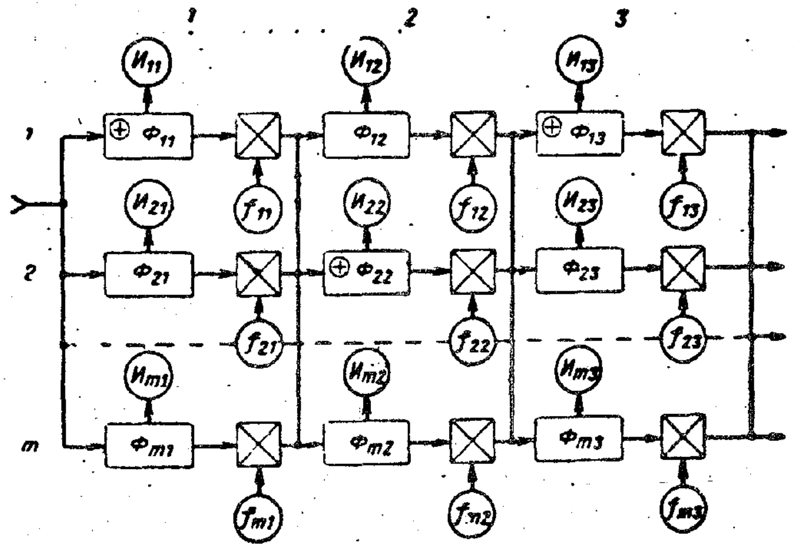


Fig. 10.22. Block diagram of matrix multichannel receiver.

The entire assigned range of reconnoitered frequencies Δf_p is split into m sub-ranges with a band Δf_1, so that

$$\Delta f_1 = \frac{\Delta f_p}{m}.$$

Frequencies of tuning of filters are shifted relative to one another by a passband. Filters of the first column (Φ_{11}, Φ_{21}, ..., Φ_{m1}) cover the entire assigned range of reconnoitered frequencies. Passband of each of these filters approximately identical and is equal to

$$\Delta f_1 = \frac{\Delta f_{\text{р}}}{m}.$$

In every column there are m heterodynes, the frequencies of which f_{11}, f_{21}, ..., f_{m1} are selected in such a way as to ensure transformation of frequencies of signals on the output of each filter to a value of intermediate frequency $f_{\text{пр}1}$ which is identical for all filters of the first column with an accuracy up to a passband of one filter of the first column Δf_1. In accordance with what was said frequencies of heterodynes of the first column are selected from the following conditions:

first line $f_{11} = f_1 + f_{\text{пр}1}$, where f_1 – lower frequency of reconnoitered range;

second line $f_{21} = f_{11} + \Delta f_1$,
third line $f_{31} = f_{21} + \Delta f_1$,...
etc.

Thus, the frequency range $f_1 - f_1 + \Delta f_{\text{р}}$ will be converted into a narrower range $f_{\text{пр}1} \div f_{\text{пр}1} + \Delta f_1$ (Fig. 10.23). The second column transforms this range into a still narrower range

$$f_{\text{пр}2} \div f_{\text{пр}} + \Delta f_2,$$

where

$$\Delta f_2 = \frac{\Delta f_1}{m} = \frac{\Delta f_{\text{р}}}{m^2}.$$

Filters of the second column have the following band of transmittance:

$$\begin{aligned}
&\Phi_{12} \to f_{\text{пр}1} < f < f_{\text{пр}1} + \Delta f_2,\\
&\Phi_{22} \to f_{\text{пр}1} + \Delta f_2 < f < f_{\text{пр}1} + 2\Delta f_2,\\
&\cdots\cdots\cdots\cdots\cdots\cdots\\
&\Phi_{m2} \to f_{\text{пр}1} + (m-1)\Delta f_2 < f < f_{\text{пр}1} + m\Delta f_2.
\end{aligned}$$

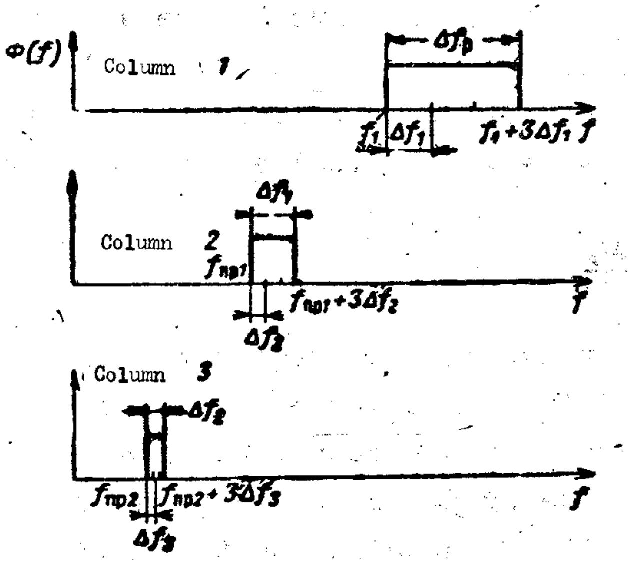


Fig. 10.23. Conversion of spectrum of signal in a matrix receiver.

Here f_{np1} — lower intermediate frequency; Δf_2 — band of transmittance of filters Φ_{12}, Φ_{22}, ..., Φ_{m2} second column.

If there are n columns, then

$$\Delta f_n = \frac{\Delta f_p}{m^n}.$$

In a general case the bands of transmittance of filters will form a unique matrix (Fig. 10.24), with help of which one can determine the frequency of the actuating signal.

No. of row	Number of column				
	1	2	3	...	n
1	$f_1 + f_1 + \Delta f_1$	$f_{np1} + f_{np1} + \Delta f_2$	$f_{np2} + f_{np2} + \Delta f_3$	...	$f_{np\,n-1} \div f_{np\,n-1} + \Delta f_n$
2	$f_1 + \Delta f_1 + f_1 + 2\Delta f_1$	$f_{np1} + \Delta f_2 +$ $+ f_{np1} + 2\Delta f_2$	$f_{np2} + \Delta f_3 +$ $+ f_{np2} + 2\Delta f_3$	...	$f_{np\,n-1} + \Delta f_n +$ $+ f_{np\,n-1} + 2\Delta f_n$
3	$f_1 + 2\Delta f_1 +$ $\div f_1 + 3\Delta f_1$	$f_{np1} + 2\Delta f_2 +$ $+ f_{np1} + 3\Delta f_2$	$f_{np2} + 2\Delta f_3 +$ $+ f_{np2} + 3\Delta f_3$	...	$f_{np\,n-1} + 2\Delta f_n +$ $+ f_{np\,n-1} + 3\Delta f_n$
⋮	...	...	...	...	...
m	$f_1 + (m-1)\Delta f_1 +$ $+ f_1 + m\Delta f_1$	$f_{np1} + (m-1)\Delta f_2 +$ $\div f_{np1} + m\Delta f_2$	$f_{np2} + (m-1)\Delta f_3 +$ $+ f_{np2} + m\Delta f_3$	...	$f_{np\,n-1} + (m-1)\Delta f_n$ $+ f_{np\,n-1} + m\Delta f_n$

Fig. 10.24. Determination of frequency of a signal with the help of a matrix receiver.

Let us assume for example, that indicators $И_{11}$, $И_{22}$ and $И_{13}$ were triggered. This means that the incoming signal passed through filters $Ф_{11}$, $Ф_{22}$, and $Ф_{13}$. The triggering of indicator $И_{11}$ (Figs. 10.22 and 10.24) indicates that the frequency of the signal f_c lies within the limits

$$f_1 \leqslant f_c \leqslant f_1 + \Delta f_1.$$

Triggering of indicator $И_{22}$ gives

$$f_1 + \Delta f_2 \leqslant f_c < f_1 + 2\Delta f_2.$$

On the force of triggering of indicator $И_{13}$ we obtain a final appraisal of carrier frequency

$$f_1 + \Delta f_2 \leqslant f_c \leqslant f_1 + \Delta f_2 + \Delta f_3.$$

Accuracy of measurement of frequency in this receiver is determined by the band of transmittance of filters of the third column Δf_3. In a general case for m lines and n columns the accuracy of determination of frequency will be equal to

$$\delta f_{\text{макс}} = \frac{\Delta f_n}{2} = \frac{\Delta f_p}{2m^n}, \tag{10.31}$$

i.e., is determined by the transmission band of the filter of the last column.

One of the main parameters of multichannel receivers is the volume of equipment, estimated in this case by the quantity of selective filters N.

For a multichannel receiver with independent channels (Fig. 10.21) the number of selective filters equals

$$N_1 = \frac{\Delta f_p}{\Delta f}. \tag{10.32}$$

For a multichannel matrix receiver (Fig. 10.22), ensuring the same accuracy of determination of carrier frequency,

$$N_2 = mn, \tag{10.33}$$

where

$$\Delta f = \Delta f_n.$$

From (10.31) and (10.33) it follows that

$$N_2 = n \sqrt[n]{\frac{\Delta f_p}{\Delta f_n}}. \tag{10.34}$$

Comparing (10.32) and (10.34), we find a gain in the number of selective filters due to the use of a matrix receiver

$$\xi = \frac{N_1}{N_2} = \frac{\Delta f_p}{\Delta f_n} \frac{1}{n \sqrt[n]{\frac{\Delta f_p}{\Delta f_n}}} = \frac{1}{n} \left(\frac{\Delta f_p}{\Delta f_n} \right)^{\frac{n-1}{n}}.$$

For example, if the number of columns $n = 3$, and $\Delta f_p / \Delta f_n = 1000$, then the gain in the number of filters is estimated by the number $\xi = 33$.

Matrix receivers are quite complex devices. The greatest difficulties in the development and tuning of multichannel matrix receivers can arise from the mutual influence between channels which generates ambiguity to measurements. This difficulty is surmounted with help of special circuits for removal of the ambiguity, separating the needed signals from interferences, and also by improvement of band filters and application of special isolating networks.

A matrix receiver ensures the best sensitivity and resolving power for frequency as compared to the standard multichannel receivers. However, time of reconnaissance (time of service) for such a receiver is somewhat greater than for an ordinary multichannel receiver.

10.6. Storage of Carrier Frequency

The goal of electronic reconnaissance may not be limited only to the determination of value of carrier frequency of reconnoitered electronic devices, but can also include memorization of this frequency, for example, in order to ensure the possibility of creation of active interferences.

Block diagrams of jamming stations with devices for storage of frequency were given earlier (Figs. 2.14 and 2.23).

In the simplest jamming stations memorization of carrier frequency and tuning the jamming transmitter on it are carried out by an operator. In this case the process of memorization of frequency and guiding (tuning) the jamming transmitter requires a great deal of time. At present there is the possibility of tuning jamming transmitters to the carrier frequency automatically.

Quantitative characteristics of different methods and devices for the storage of frequency are the following:

- time for tuning;
- storage time (memory);
- accuracy of tuning;
- accuracy of retention of frequency;
- range of memorization;
- resolving power (ability of simultaneous tuning on several frequencies).

Let us consider certain concrete methods and devices for the storage of frequency.

Memorization of Carrier Frequency with Help of Automatic Tuning of Generator (Method of Automatic Frequency Control APCh for Deflection)

This method uses the well-known principle of automatic frequency control of heterodynes of receivers which is applied extensively at present. The block diagram of a single-channel device for memorization of frequency is shown in Fig. 10.25. The signal of the suppressed radio electronic device enters the amplifier y, through

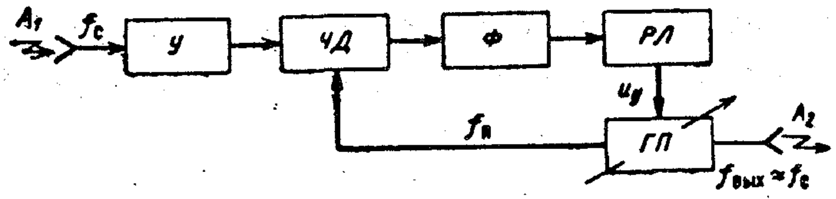


Fig. 10.25. Block diagram of single-channel device of memorization of frequency with help of automatic tuning of generator.

the receiving antenna A_1, after which it acts on the frequency discriminator ChD. Also conveyed here is the voltage of the tuned jamming generator GP. In the event of deviation of frequency of jamming generator $f_{п}$ from the frequency of the actuating signal f_c at the output of the frequency discriminator there appears a voltage, which after filtration exerts an influence on the reactance tube RL, with the help of which control over the frequency of the jamming generator is carried out. The circuit is built in such a way that in the event of the appearance of a mismatch the controlling action brought it to zero. Thus the frequency of the jamming generator is maintained close to the carrier frequency of the suppressed electronic device.

The described circuit requires considerable uncoupling of receiving A_1 and transmitting A_2 antennas. In flying jamming stations the uncoupling of receiving and transmitting antennas encounters significant difficulties.

Deficiencies of this single-channel device are:

— small width of range of memorization, limited by possibilities of circuits of electronic and mechanical frequency control;

— insufficient resolving power (the circuit memorizes only one frequency).

Memorization of Carrier Frequency by Means of Automatic Fine Adjustment of Generator by Extreme of Signal (Method of Automatic Frequency Control Based on Extreme)

The fundamental possibility of application of methods of extreme adjustment for purposes of memorization of frequency is conditioned by the relative simplicity of obtaining an extreme of signal with the help of elements which possess frequency selectivity (filters, resonant circuits, and so forth), and also realization of scanning devices, for example, by means of electronic retuning of the heterodyne or electromechanical retuning of the oscillatory circuit [94].

Figure 10.26 shows the circuit of a frequency storage device [95], built on the principle of automatic frequency control based on extreme. The signal of the suppressed electronic device with carrier frequency f_c after amplification in the wideband amplifier ShU proceeds to the oscillatory circuit, consisting of variable capacitors C_1, and C_2 variable inductance L_1.

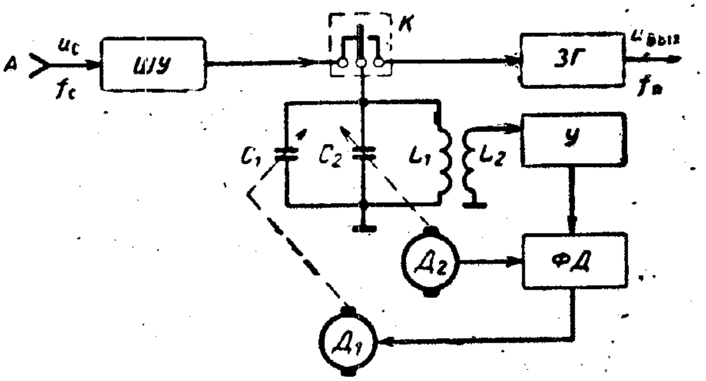


Fig. 10.26. Block diagram of extreme single-channel device for frequency storage.

Capacitor C_2 has a comparatively small capacitance, the value of which is changed periodically with the help of motor Д_2. Due to this the compulsory small changes of resonance frequency of the circuit are induced. These are necessary for searching for the extreme in the established operating conditions of the system.

Periodically variable voltage, corresponding to the change of capacitance of capacitor C_2, moves from motor to one of the inputs of the phase detector Д_2, on the second input of which amplified oscillation arrives from the oscillatory circuit. The phase detector forms a signal of deflection of resonance frequency of the circuit from extreme which is fed to the motor Д_1, turning the rotor of the main capacitor C_1.

At the moment of termination of tuning of the oscillatory circuit in resonance the commutator K disconnects it from the input circuit and connects it to the corresponding circuits of the master oscillator ZG, passes into a condition of automatic oscillations on a frequency close to frequency of the actuating signal $(f_\text{п} \approx f_\text{c})$.

We will examine the process of searching for an extreme in a particular device. The capacitor for forced scanning C_2 produces a periodic oscillation of resonance frequency (f_p) of the circuit, which is equivalent to the periodic frequency shift (f_c) of the input signal during constant tuning of the circuit (Fig. 10.27). The periodic change of resonance frequency causes modulation of voltage on the output of the oscillatory circuit (curves 1, 2, 3, Fig. 10.27). Amplitude and phase (and frequency) of this voltage depend on the value and sign of detuning of the circuit relative to frequency of the actuating signal. With an exact coincidence of resonance frequency of the circuit (average for the period of scanning) with the frequency of the actuating signal ($f_c = \overline{f}_p$) the frequency of the first harmonic of the envelope of output voltage is equal to double the frequency of scanning (curve 1, Fig. 10.27). If, however, the average value of resonance frequency for the period of investigation differs from the frequency of the actuating signal, then the frequency of the first harmonic of output voltage is equal to the frequency of search for the extreme and its phase and amplitude correspond to the sign and value of detuning of the circuit relative to frequency of external signal (curves 2, 3, Fig. 10.27).

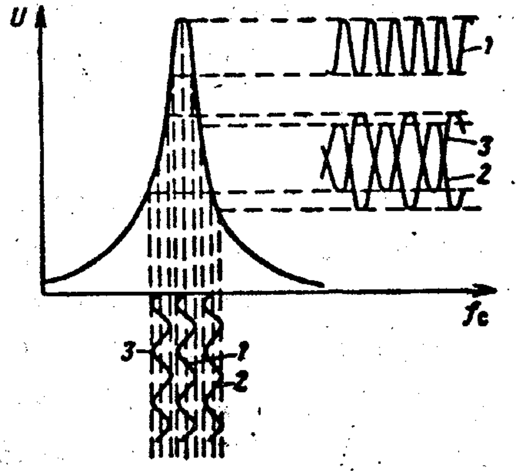


Fig. 10.27. Scanning for the extreme in a single-channel device for frequency storage.

At the output of the phase detector, which is carrying out the operation of multiplication and averaging of signals arriving on its inputs, in the first case voltage will be equal to zero

$$\frac{1}{T}\int_0^T \cos(\omega t+\varphi)\cos 2\omega t\,dt = 0,$$

and in the second case it is proportional to cos ϕ:

$$\frac{1}{T}\int_0^T \cos(\omega t+\varphi)\cos \omega t\,dt \sim \cos\varphi.$$

Thus, in the event of error in tuning of the resonance circuit the phase detector produces a mismatch signal which controls the motor Д_1. The latter turns the rotor of the variable capacitor until the average resonance frequency of the circuit coincides with the frequency of the output signal (accordingly on the output of the FD the voltage will be equal to zero). This arrangement can be carried out comparatively simply in the meter and decimeter ranges of waves.

Multichannel Method of Frequency Storage

The examined method of memorization is a development of the multichannel method of reconnaissance of frequency (Fig. 10.21). The range of memorization is covered by a system of filters. After amplification and detection voltage from the output of these filters proceeds to the relay $\text{Рл}_1, \text{Рл}_2, \ldots, \text{Рл}_m$.

If in some channel j a signal is revealed, then relay Рл_j is triggered and the corresponding jamming enerator Γ_j is switched on (Fig. 10.28). Accuracy of memorization of frequency with such method is determined by bandwidth of transmission of the input filters.

A basic deficiency of this arrangement is considerable volume of equipment, if one is concerned with ensuring memorization with high accuracy in a wide frequency range.

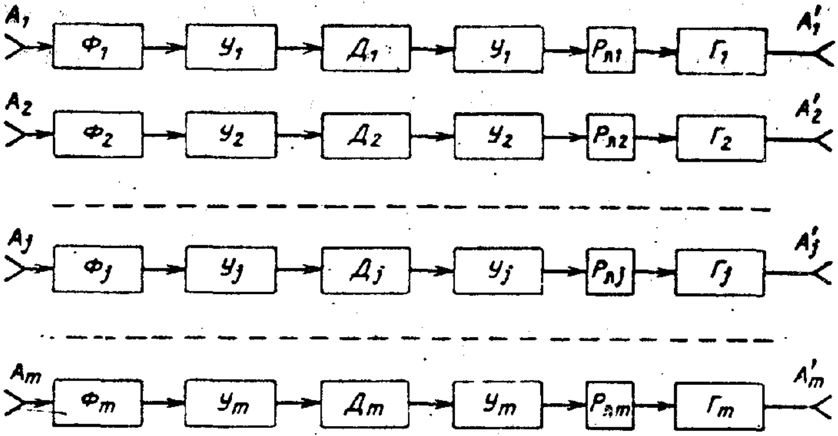


Fig. 10.28. Block diagram of multichannel device for frequency storage.

10.7. Direction Finding of Electronic Devices in the Interests of Electronic Reconnaissance

Knowledge of angular coordinates of electronic devices makes it possible to determine their position and when necessary to guide the antennas of jamming transmitters on them.

Direction finding devices of electronic reconnaissance stations should satisfy the following basic requirements:

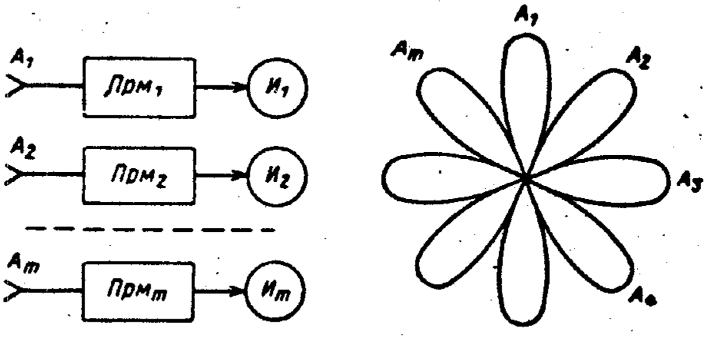
- ensure measurement of bearing in the shortest possible time;
- have a sufficiently high accuracy and resolving power for angular coordinates in a wide frequency range.

In electronic reconnaissance nonscanning and scanning methods of direction finding of sources of emission are used.

Nonscanning methods of direction finding make it possible to determine direction to the source of emission instantly regardless of the location of the source relative to the antenna of the direction finder (within the range limits of electronic reconnaissance). Scanning methods of direction finding make it possible to determine direction to the source by means of the systematic examination of the reconnoitered space. Determination of the bearing of a source of emission in this case requires a certain amount of time. Both these methods of direction determination can use all forms of radio direction finding: amplitude, phase, and frequency. Amplitude and phase radio direction finders are used most extensively.

Nonscanning Methods of Direction Finding

In the simplest case nonscanning determination of direction to a source can be carried out with help of a multichannel space-selective device. A block diagram of such an arrangement, intended for determination of direction in one plane, is shown in Fig. 10.29a. Reception of signals is carried out by antennas (A_1, A_2, ..., A_m) from all directions. The directional pattern of antennas is depicted in Fig. 10.29b. Accuracy of determination of direction and resolving power here are determined by half the width of the antenna directional pattern on a level of 0.1-0.05.



a) b)

Fig. 10.29. Nonscanning method of determination of bearing of an electronic device: a) block diagram of multichannel space-selective device; b) total antenna directional pattern of receiving device.

A high degree of accuracy in the determination of bearing can be ensured with help of a large number of antennas and, consequently, receiving channels. This is an essential deficiency of the described circuit.

Good characteristics belong to the so-called functional direction finder device, the principle of operation of which is based on the functional dependence of output total voltage of two or several antennas on the direction of arrival of radio waves.

An example of functional direction finder devices can be the automatic radio direction finders, which are well-known from radio navigation, with an H-shaped antenna array, and also the automatic radio compasses which use a frame radio direction finder for determination of direction.

Functional direction finders have a high degree of accuracy for determination of direction and make it possible to orient two radio transmitters which are working on the same frequency. They have been used successfully in the range of meter waves in various ground electronic devices. Their practical application on aircraft in the meter range of waves runs into serious difficulties connected with the dimensions of antennas and especially with the nonuniform influence of the airplane fuselage on antenna radiation pattern based on frequency range.

Scanning Methods of Direction Finding

These methods found wide application in aircraft stations for electronic reconnaissance. Determination of direction to the source of emission is carried out with the help of a revolving high-directional antenna combined with an electron-beam indicator, in which the scanning line shifts synchronously with rotation of the antenna, thus forming a coordinate scale. The blip of the incoming signal can be amplitude or based on brightness. Usually direction finding is conducted by the method of maximum. The bearing on the radio transmitter in this case is determined by the angular position of the high-directional antenna at which the signal of the reconnoitered electronic device on the output of the direction finder reaches maximum value.

Let us consider the peculiarities of determination of bearing on radar which is working under conditions of scanning (Fig. 10.30). With the circular rotation of the radar antenna with angular

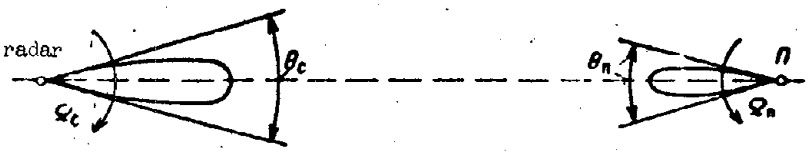


Fig. 10.30. Scanning method for determination of bearing.

frequency Ω_c at the point of reception (П) we obtain a series of pulses, following with a frequency of $F_c = \Omega_c/2\pi$. Duration of series τ_c is characterized by width of antenna radiation pattern of the radar θ_c, i.e.,

$$\tau_c = \frac{\theta_c}{2\pi F_c} = \frac{\theta_c}{\Omega_c}.$$

In Fig. 10.31 is depicted aximuthal-time diagram of scanning, constructed in rectangular coordinates. Time segments, during which signals from the radar is possible, are noted on the figure by heavy lines. As can be seen from the figure, in general the determination of direction takes place with a certain probability.

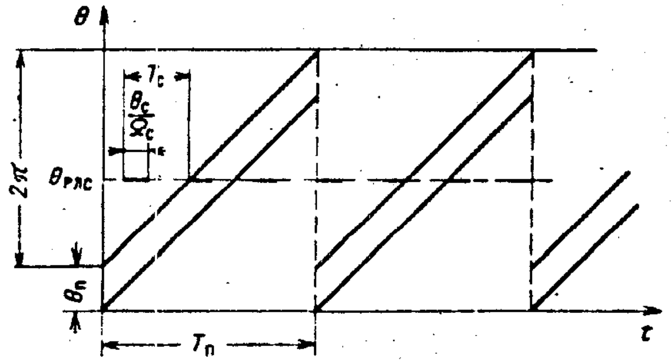


Fig. 10.31. Azimuthal-time diagram for determination of bearing of radar.

Probability of determination of direction depends on width of beams of radar antennas (θ_c) and the electronic reconnaissance station ($\theta_п$).

For the assigned system of electronic reconnaissance it can be determined on the basis of the queueing theory which was expounded above or the theory of coincidences. Below is expounded a simplified method for appraisal of probability of determination of direction which is useful in estimated calculations [23].

If the angular velocity of rotation of the scanning high-directional antenna of the direction finder is great in comparison to the angular velocity of rotation of scanning antenna of the radar $(\Omega_{\text{п}} \gg \Omega_{\text{c}})$, then the probability of interception of the signal during the time of one revolution of the radar antenna is equal to the ratio of exposure time of the radar in a given direction for one period of rotation of the antenna $\frac{\theta_c}{\Omega_c} = \frac{\theta_c T_c}{2\pi}$ to the duration of one period of rotation of the antenna of the direction finder $T_{\text{п}}$:

$$P_1 = \frac{\theta_c T_c}{T_{\text{п}} 2\pi}.$$

The probability of interception of the signal during two rotations of the antenna of the direction finder will equal

$$P_2 = P_1 + (1 - P_1) P_1,$$

and for n rotations

$$P_n = P_1 + (1 - P_1) P_1 + (1 - P_1) P_1^2 + \ldots + (1 - P_1)^n P_1.$$

Summing up the resulting geometric progression, we obtain

$$P_n = 1 - (1 - P_1)^n.$$

The last expression can be rewritten in the following way:

$$P_n = 1 - \left[(1 - P_1)^{-\frac{1}{P_1}} \right]^{-nP_1}.$$

If probability of interception of signal (P_1) during one revolution of the antenna of the direction finder is small, then, considering that

$$\lim_{P_1 \to 0} (1 - P_1)^{-\frac{1}{P_1}} = e,$$

we obtain for the resultant probability of detection of the radar signal in n cycles of scanning of the high-directional antenna of the direction finder the following formula:

$$P_n = 1 - e^{-nP_1}.$$

Since the number of scanning cycles (n) is equal to the overall time of scanning t_p, divided by period of scanning T_c:

$$n=\frac{t_p}{T_c},$$

then the final expression for the probability of detection of the signal during the time t_p will have the form

$$P_n=1-e^{-\frac{\theta_c t_p}{2\pi T_п}}. \tag{10.35}$$

Formula (10.35) shows that the probability of detection of the signal aspires to a unit with an increase of general time of reconnaissance t_p and a decrease in the period of scanning $T_п$.

Rate of rotation of the antenna of the direction finder cannot be infinitely great. It is limited by the time, necessary for reception of at least one pulse of radar, in other words, the time, during which the antenna of the direction finder can receive signals from a given direction, should be no less than the period of tracking of pulses of the reconnoitered radar

$$\frac{T_п\theta_п}{2\pi}\geqslant T_и=\frac{1}{F_и}.$$

It follows from this that the scanning antenna should have as wide an antenna radiation pattern as possible. However, the increase of $\theta_п$ is limited by requirements of accuracy and resolving capacity.

Just as during determination of frequency, slow and rapid scanning of signal for direction are possible.

During slow scanning the rate of rotation of the antenna of the direction finder is selected as such, so that during the time $\Delta T_п$ of passage (by the antenna of the direction finder) of an angle which is equal to the width of the major lobe of its own antenna radiation pattern, the antenna of the radar made at least one revolution, i.e.,

$$\Delta T_п\leqslant T_c=\frac{2\pi}{\Omega_c}.$$

During rapid scanning the rate of rotation of the antenna of the direction finder is selected as such, so that during the time (ΔT_c) of passage (by the antenna of the radar) of an angle which is equal to the width of the major lobe of its own antenna radiation pattern, the antenna of the direction finder made at least one turn, i.e.,

$$\Delta T_c \leqslant T_{\text{п}} = \frac{2\pi}{\Omega_{\text{п}}}.$$

In practice this condition is far from always feasible. Most frequently scanning for direction is ensured with a certain average speed. In a general case the necessary rate of scanning is determined by methods of the queueing theory and the theory of coincidences of pulse flows on the basis of permissible value of probability of electronic reconnaissance.

10.8. Determination of Position of Electronic Devices

The position of electronic devices of the enemy can be determined both by direct and indirect methods. By direct methods is understood the measurement of position of source as a result of direct processing of incoming signals. In indirect methods the determination of position of a source is carried out by formulas, connecting the coordinates of source with its bearings as derviced from several points, and the distances between points of measurement of bearings.

Direct Methods of Determination of Position of Sources of Emission

An example of this method can be the so-called vertical method of surveying space which is applied during electronic reconnaissance with the help of artificial earth satellites (ISZ). This method requires the flight of an artificial earth satellite over the reconnoitered electronic device (Fig. 10.32). Interception of signals is carried out by a narrow-directional antenna. At the time of interception of signals a record is made of the position of the point of interception.

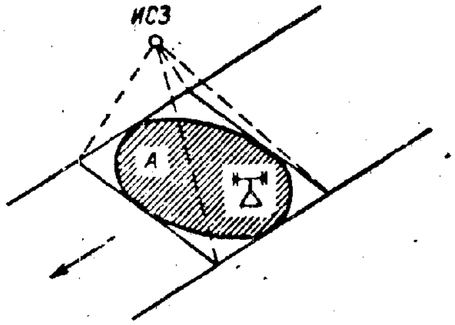


Fig. 10.32. Region of uncertainty of geographic position of an electronic device during the direct method of determination of position.

The region of uncertainty of the geographic position of a detected source of emission is characterized by the so-called geographic resolving power of the system, which is determined by area (A) of a region which is surveyed simultaneously by the receiving antenna of an electronic reconnaissance station. For an antenna radiation pattern which is circular in a cross section the geographic resolving power is determined by the relationship

$$A = \pi h^2 \mathrm{tg}^2 \frac{\theta}{2},$$

where h – altitude of flight; θ – aperture angle of directional pattern.

With a single flight the accuracy of determination of position of source is small. It can be increased at the expense of multiplicity of surveying of the assigned space with a mutual covering of areas embraced by the receiving antenna during each flight.

Indirect Methods of Determination of Position of Sources of Emission

The most wide-spread is direction finding of a source of emission from two or more points located on a known base line, with the subsequent calculation of its position by the method of triangulation (Fig. 10.33). Such a method of determination of position creates an area of uncertainty (A) at the point of intersection of the radiation patterns of the receiving antenna. It is possible to show that the best geographic resolving power will be obtained, if the moments of interception correspond to bearings on the source of $\theta_1 = \theta_2 = 60°$

$$A = A_{\text{мин}} = 6{,}2 R^2 \mathrm{tg}^2 \frac{\theta_{0,5}}{2},$$

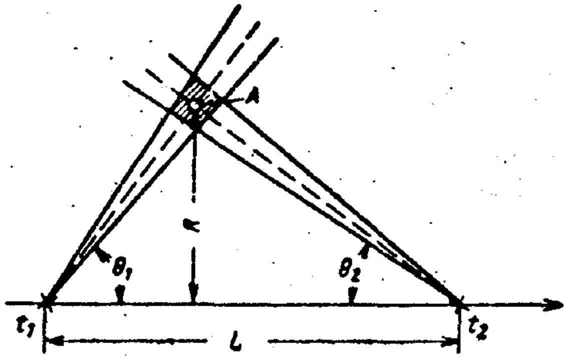


Fig. 10.33. Region of uncertainty of geographic position of an electronic device during the indirect method of determination of position.

where R – minimum distance between line of flight and source of emission; $\theta_{0.5}$ – width of beam of receiving antenna of direction finder.

For realization of two interceptions of signals which are successive in time either one rotating antenna is used or two identical antennas, set up at an angle of 60° to the axis of the aircraft.

In conclusion we will note the basic differences between direct and indirect methods of determination of position. In the first case determination of position in principle can be carried out at the expense of directed reception of a signal in one point, whereas in the indirect method it is necessary to carry out reception and direction finding in a minimum of two points of space.

10.9. Peculiarities of Identification of Form of Electronic Device

Identification of form requires the fulfillment of two basic operations (10.2): formation of vector of criteria on the basis of a posteriori information and a comparison of it with the vector of criteria, formed on the basis of a priori data.

Identification of form of electronic devices and determination of their tactical and technical data (appraisal of parameters) are conducted on the basis of an analysis of detected radio emissions. The analysis of incoming radio signals includes a series of successive operations, the main ones of which are:

- recording of signals;
- measurement of parameters of signals;
- processing of information obtained.

All these operations are fulfilled mainly by the operator and partially automatically and semiautomatically by analyzers and recorders. Identification of form of a reconnoitered electronic device and determination of its tactical and technical data (appraisal of parameters) is carried out on the basis of a comparison of parameters measured in the process of electronic reconnaissance with a priori known parameters of signals of electronic devices of the enemy.

As criteria for an electronic device one can accept first of all all four independent parameters of any signal: amplitude, frequency, phase, polarization. Each of these parameters can be changed in time either purposely or accidentally. Accordingly the criteria can be parameters of this or that modulation, parameters of laws of distribution of random variables.

10.10 Principles of Control of Jamming Devices

The final goals of electronic reconnaissance are:

- determination of the most dangerous electronic devices of enemy antiaircraft defense, the suppression of which must be carried out immediately;
- determination of a rational form of jamming signal and operating conditions for the corresponding jamming transmitter.

In the determination of form of jamming signal and operating conditions for the jamming transmitter it is necessary to consider that jamming not only inflicts information loss on the enemy, but also gives to him certain new possibilities for actions. By disrupting the operation of some means of control of the enemy, jamming can facilitate the application of other principles. For example, setting out strips of chaff lowers probability of damage to covered targets, but at the same time probability of detection of the entire group of aircraft on the whole is increased. Active jamming under definite circumstances can facilitate the enemy's detection of the target and guiding of means of destruction on it.

In the process of overcoming enemy antiaircraft there is great significance in the selection of the moment for beginning the operation of jamming facilities (moment of switching on of jamming transmitters, automatic machines for dropping reflectors, launching of traps, and so forth). If jamming devices are switched on too early, then this increases the distance at which enemy antiaircraft defense detects the target, and is able, by turning to cope with the given interference, to organize counteractions. If, however, they are switched on too late, then the antiaircraft defense can be in a state to use its own capabilities and destroy the aircraft still prior to the beginning of operation of radio countermeasures. Thus, an optimum time exists for switching on of radio countermeasures and an optimum duration for their operation.

Different contours of antiaircraft defense have a different sensitivity of their effectiveness relative to the moment of switching on of radio countermeasures and duration of their operation. Figure 10.34 depicts the qualitative dependence of effectiveness of the contour of target distribution on the range of switching on of radio countermeasures. By effectiveness of the contour of target distribution can be understood the number of successful attacks, number of valleys on attacking aircraft, and so forth. In this case along the axis of ordinates is plotted the probability of destroying the target $P_{сб}$.

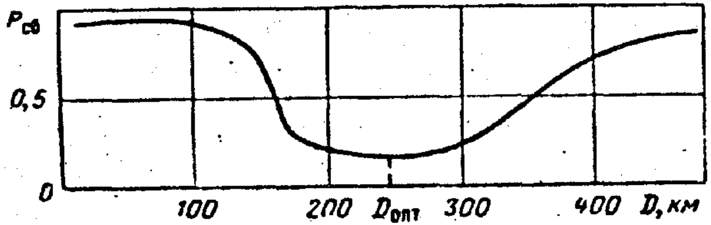


Fig. 10.34. Qualitative dependence of effectiveness of antiaircraft defense of an object (contour of target distribution) on the distance of switching on of radio countermeasures.

Counteraction against the electronic devices of contours of target distribution can be carried out by different radio countermeasures. Moments of switching on and duration of their operation have to be determined beforehand on the basis of data, obtained with the help of preliminary electronic reconnaissance, and as soon as possible be definitized in the process of overcoming the antiaircraft defense.

Optimum time for beginning the jamming of electronic devices of systems for laying and homing guidance can be determined by the moment of launching of rockets or switching over of on-board radar of fighter aircraft, the target tracking radar of antiaircraft guided missiles, and homing devices of guided missiles to conditions of automatic tracking.

If the covered aircraft has an appropriate complex of reconnaissance equipment, making it possible to detect and measure parameters of emitted signals, then with its help it is possible to establish:

- the presence of continuous automatic tracking of the aircraft by illumination radar of the target or the on-board radar of a fighter (rocket);
- direction of attack, and also other necessary data.

There is interest in the use of an electronic digital computer for a posteriori formation of vector of criteria of electronic devices. Let us note that, leaning on the contemporary level of computer development, in the on-board complex of electronic reconnaissance and control of jamming signals one should be oriented on them mainly as a means of processing and generalization of reconnaissance information. Responsible decisions relative to radio countermeasures, even on board an aircraft, apparently should be made by a person who has been appropriately trained for this.

We will examine certain general principles of controlling jamming devices and signals in the process of overcoming antiaircraft defense on the basis of information criteria.

The system of antiaircraft defense on the whole can be examined as a complex system of a base type – a Б-system [97].

Functioning of Б-systems in general is investigated by methods of the queueing theory. Elements of the Б-system are pickups of information (radar), points for collection and processing of data and making of decisions (control), means of transmission of information, command elements, and so forth. Elements of the Б-system are separate closed weapon – control systems constituting the simplest system of automatic adjustment – a self-guided missile, a system of automatic tracking for direction, and others.

Thus, we will relate to simple systems all contemporary devices for automatic adjustment. At the same time a complex of simple automatic devices controlled by man, for example a complex of guided antiaircraft missiles with target designation radar, by a device for analysis of jamming signals, by the commander, and by operator-plotters, we will relate to complicated system.

In the functioning of all the most important sections of a system of antiaircraft defense it is possible to distinguish two basic stages.

In the first stage an appraisal of the air situation is made and based on its results a decision on subsequent actions is made. Here the main element is resolution of the problem of target distribution.

In the second stage the accepted solution is realized, i.e., at first turn the guiding of the aircraft (guided antiaircraft missiles) is carried out and also other measures are conducted (notification, rebasing, camouflage, and so forth).

The system of antiaircraft defense, just as any other Б-system, functions on the basis of information arriving from the corresponding pickups. In the system of antiaircraft defense the main pickups of information are radar for detection, guiding, and target designation, and also radar for the control of weapons.

Information about the air situation which is received from the radar system removes the a priori uncertainty and makes it possible in the very best manner to make a decision corresponding to the situation which is forming at the given moment of time.

The multiplicity of possible states of a complex base system is finite, although the overall number of states can be quite large.

Selection of this or that state of the Б-system is determined by the number and type of demands of objects, called customers, which are serviced by the system. In the case of the antiaircraft defense system the customers are aircraft and winged rockets of attacking aviation.

Aircraft and winged rockets of attacking aviation, in a certain sense, also constitute a complex control system, inasmuch as in this system in principle a circulation of information can take place, for example, ensuring control of aircraft in combat formation or control of jamming transmitters which are mounted on these aircraft.

If one were to be limited only to control of jamming devices on the basis of data arriving from aircraft and other electronic reconnaissance equipment which is found in the air, then attacking aviation can also be referred to class Б-systems.

One or several states of attacking aviation corresponds to one or several states of the system of antiaircraft defense. Selection of corresponding state of antiaircraft defense is made on the basis of information obtained about the state of the attacking system. Subsequently the system of antiaircraft defense will be designated the $Б_{П}$-system, and the system of attacking aviation – $Б_{A}$-system.

Information of the $Б_{A}$-system does not remove completely the a priori uncertainty. A certain a posteriori uncertainty (entropy) remains in the $Б_{П}$-system. For this reason the probability of making the optimum decision, i.e., selection of the state of the $Б_{П}$-system, in the very best manner corresponding in the accepted criterion to the state of the $Б_{A}$-system, will not be equal to a unit. In other words, on the force of a posteriori uncertainty the following finite probability scheme takes place for the $Б_{\Pi}$-system at the time of making the decision:

$$B = \begin{pmatrix} B_1 B_2 \dots B_i \dots B_n \\ P_1 P_2 \dots P_i \dots P_n \end{pmatrix}. \qquad (10.36)$$

Here B_i – one of the possible states of the system ($i = 1, 2, \dots, n$);

P_i – probability to find the system in state B_i after acceptance of a decision with the given a posteriori uncertainty.

Thus, for the great number $\{B_n\}$ of possible states of the $Б_{\Pi}$-system at a given moment of time with an assigned a posteriori uncertainty probability distribution P_i, takes place, where not all $P_i = 0$ and $\sum_{i=1}^{n} P_i = 1$.

Each value of a posteriori uncertainty has its corresponding, in a general case different from (10.36), finite probability scheme, describing the state of uncertainty of the $Б_{\Pi}$-system at the moment of making a decision. The value of this uncertainty – entropy $H_r(P_1, \dots, P_n)$, which we will call entropy of decision, can be determined by the known formula:

$$H_r(P_1, P_2, \dots, P_n) = -\sum_{i=1}^{n} P_i \log P_i.$$

The greater the a posteriori uncertainty at the time of making the decision, then, generally speaking, the greater the entropy of the decision $H_r(P_1, \dots, P_n)$ of the $Б_{\Pi}$-system.

Consequently the $Б_A$-system should function in such a way that at the time of acceptance of a decision in $Б_{\Pi}$-system its entropy of decision is maximum. In other words, it is necessary to control these jamming devices and the entire $Б_A$-system so that acceptance of a decision in the $Б_{\Pi}$-system takes place at a maximum of a posteriori uncertainty – a maximum of a posteriori jamming taking place after partial removal by system of information safeguard of a priori uncertainty.

In general the attacking side does not know a priori at what moment of time a decision will be made, especially if the aircraft are in the depth of the enemy's antiaircraft defense. Therefore it is expedient to act in such a manner so that in every given moment of time during a certain interval a maximum of uncertainty was ensured in the system of information safeguard for the antiaircraft defense.

In view of countermeasures by the Б_{Π}-system, directed towards combatting the jamming, the value of entropy will be a random function of time.

Considering simplification of the calculation, we will present the process of change of entropy in time in the form of a multistage process. Every time interval Δt_i of a multistage process is given in conformity to the value of a posteriori entropy $H_{\text{п}i}$ and the probability of $P_{\text{п}i}$ is such that this entropy takes place. Then average entropy for a certain assigned time interval T

$$\langle H_{\text{п}} \rangle = \sum_{i=1}^{m} H_{\text{п}i} P_{\text{п}i}. \tag{10.37}$$

If density of distribution p(t) of a posteriori jamming entropy is known as a function of time, then

$$\langle H_{\text{п}} \rangle = \int_{0}^{T} p(t) H_{\text{п}}(t)\, dt.$$

Thus, as a direct criterion of optimum control of jamming devices intended for suppression of a system of target distribution, it is possible to take the average a posteriori jamming entropy in the system of radar and other information safeguard of the Б_{Π}-system.

Considered as best is control which at assigned limitations on the $\text{Б}_{\text{А}}$-system ensures a maximum of mean value of a posteriori jamming entropy in the system of information safeguard of the Б_{Π}-system.

The above-stated principle will be called the principle of maximum of jamming entropy — max $\langle H_{\text{п}} \rangle$.

It is simple to see that a conformity takes place between probability of overcoming the antiaircraft defense, mathematical expectation of losses of attacking aviation, and value of a posteriori jamming entropy in the system of information safeguard of the antiaircraft defense. It is obvious that if the attacking side uses all its possibilities taking into account assigned limitations and ensures a maximum of uncertainty in every given moment of time during assigned interval T, then this will ensure a minimum of losses in the Б_A-system and maximum of average probability of overcoming the antiaircraft defense by each aircraft.

With assigned forces and means, and also conditions of application, realization of the principle of maximum of jamming entropy ensures the least losses as compared to other methods of actions. Numerical value of losses is estimated on the basis of operational-tactical criteria.

Depending on the type of jamming signal, jamming entropy is determined differently.

For continuous noise jamming signals and passive jammings the jamming entropy can be determined by [98] as ε-entropy of a great number in metric space:

$$H_{\text{п}}(t) = \log \frac{V_{\text{п}}}{V_{\text{по}}}.$$

Here $V_{\text{п}}$ — mean value of camouflaging by jamming of the volume of space in the zone of radar observability: $V_{\text{по}}$ — mean value of pulse volume of radar or average resolving power of system in the serviced space.

In case of imitation jammings and false targets jamming entropy can be determined with help of a probabilistic circuit of the following form:

$$A = \begin{pmatrix} A_1 A_2 \ldots A_j \ldots A_l \\ q_1 q_2 \ldots q_j \ldots q_l \end{pmatrix}, \qquad (10.38)$$

where each target is characterized by a vector of features

$$T_i = \left\| \begin{matrix} \tau_1(P_1) \\ \tau_2(P_2) \\ \cdot \\ \cdot \\ \cdot \\ \tau_k(P_k) \end{matrix} \right\|.$$

Matrix-column T_i describes the vector of criteria of a true or false target. Any target is determined by k-measuring vector of criteria; each of the criteria can be assigned with a probability distribution P_{ki}.

Circuit A describes the a priori conditions of selection of true targets from a great number $\{A_i\}$. Elements of great number $\{A_i\}$ are all possible events, taking place in process of selection of true targets against the background of a certain number of false targets.

Probabilities of occurrence of these events q_j are determined by the relationship quantity of true and false targets, and also by probabilities of the presence of corresponding components for vectors of criteria of each of the targets. We recall that by definition

$$\sum_{j=1}^{l} q_j = 1.$$

In a general case of probability q_j should be determined on the basis of theory of static solutions taking into account the limitedness of time of observation of targets, which has a definite meaning. The initiator of jamming should aspire so that jamming entropy $H_n(A)$, corresponding to the probabilistic circuit (10.38), reaches the highest possible values under the given conditions.

Probabilities of P_{ni} in (10.37) permit the considering, in principle, of a response action of antiaircraft defense in the struggle with jamming.

In general, in the most general situation P_{ni} are determined by the results of a solution of game problem, corresponding to the given moment of time. However, a calculation of response actions can also be made without the solution of the game problem.

At present two basic methods of dealing with jammings are known: radio electronic and fire. By the fire method is understood a method of combat, founded on destruction of the aircraft-carriers of jamming devices. By radio electronic are understood all the remaining methods of an organizational-tactical and technical nature which are connected with the application of radio electronic equipment under conditions of jamming.

Radio electronic methods of combat, such as retuning to carrier frequency, introduction of additional, previously hidden devices, increase of power potential, and so forth in principle can be considred as the corresponding configuration of jamming equipment.

Calculation of fire counteraction is a quite complex problem if one were to talk about application of the means of radio counteraction against the system of target distribution. Therefore, in calculations of the first approximation it is possible to disregard fire opposition, assuming that after individual aircraft with jamming devices are brought down the remainder have to operate in such a way that under the new conditions at every given moment they ensure a maximum of a posteriori jamming entropy and, besides that, ensure a maximum of average jamming entropy.

In conclusion let us note that the effectiveness of radio countermeasures, just as any other means of armed combat, depends both on technical indices and on operational-tactical perfection of methods of their application. Therefore the scientifically proven treatment of tactics for the application of radio countermeasures has a decisive value.

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